JERROLD®

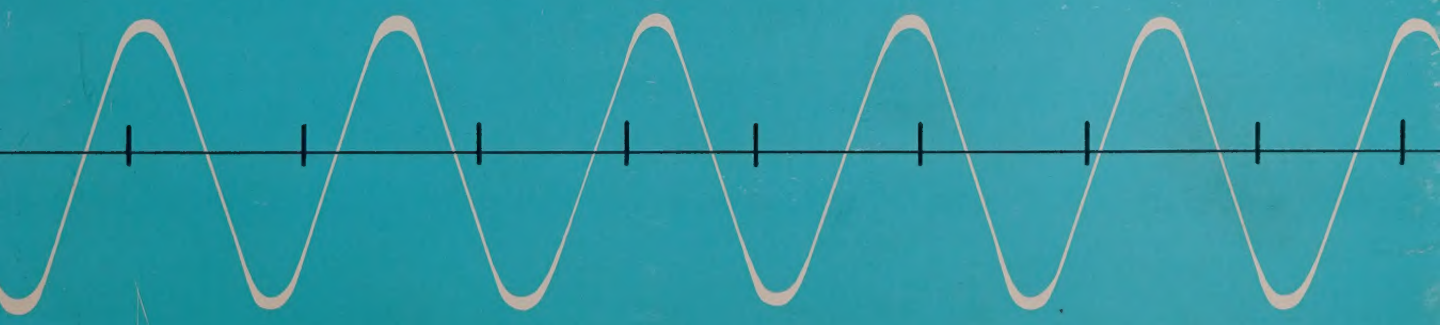
Instruction Manual

Ed Helming

# Crystal Controlled
# *Marker Generators*
## Models CM-6, CM-10

JERROLD MBC

PRECISION INSTRUMENTS FOR
MEASUREMENTS BY COMPARISON



ST. #435-275.1

# CRYSTAL CONTROLLED MARKER GENERATORS

## MODELS CM-6 AND CM-10

## Instruction Manual IP-PM-001

JERROLD® ELECTRONICS CORPORATION
INDUSTRIAL PRODUCTS DIVISION
THE JERROLD BUILDING, PHILADELPHIA 32, PA.

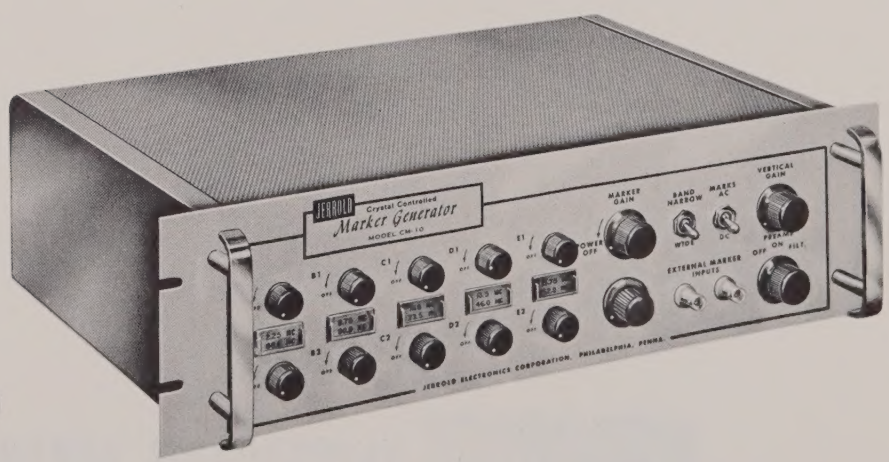
Copyright © 1961, Jerrold Electronics Corporation

All Data Subject to Change Without Notice

JERROLD

# Crystal Controlled Marker Generators

**MODEL CM-6**




**MODEL CM-10**

# TABLE OF CONTENTS


| | PAGE |
|---|---|
| General Description | 1 |
| Specifications | 1-2 |
| Chart of Controls | 3 |
| Circuit Description | 4 |
| Operation and Maintenance | 5-8 |
| Warranty | 8 |
| Parts List, CM-6 | 10-11 |
| Parts List, CM-10 | 12-13 |
| Parts List, Oscillators | 14-15 |

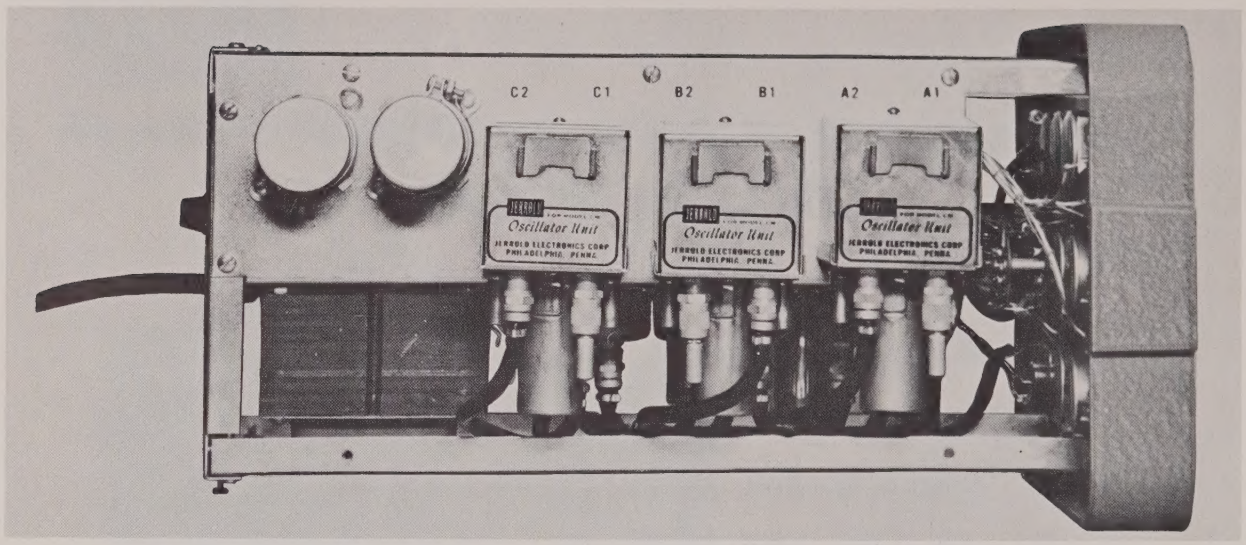



**Figure 1—Jerrold Crystal-Controlled Marker Generator—Model CM-6**
**Top view of Chassis showing oscillator plug-in units.**

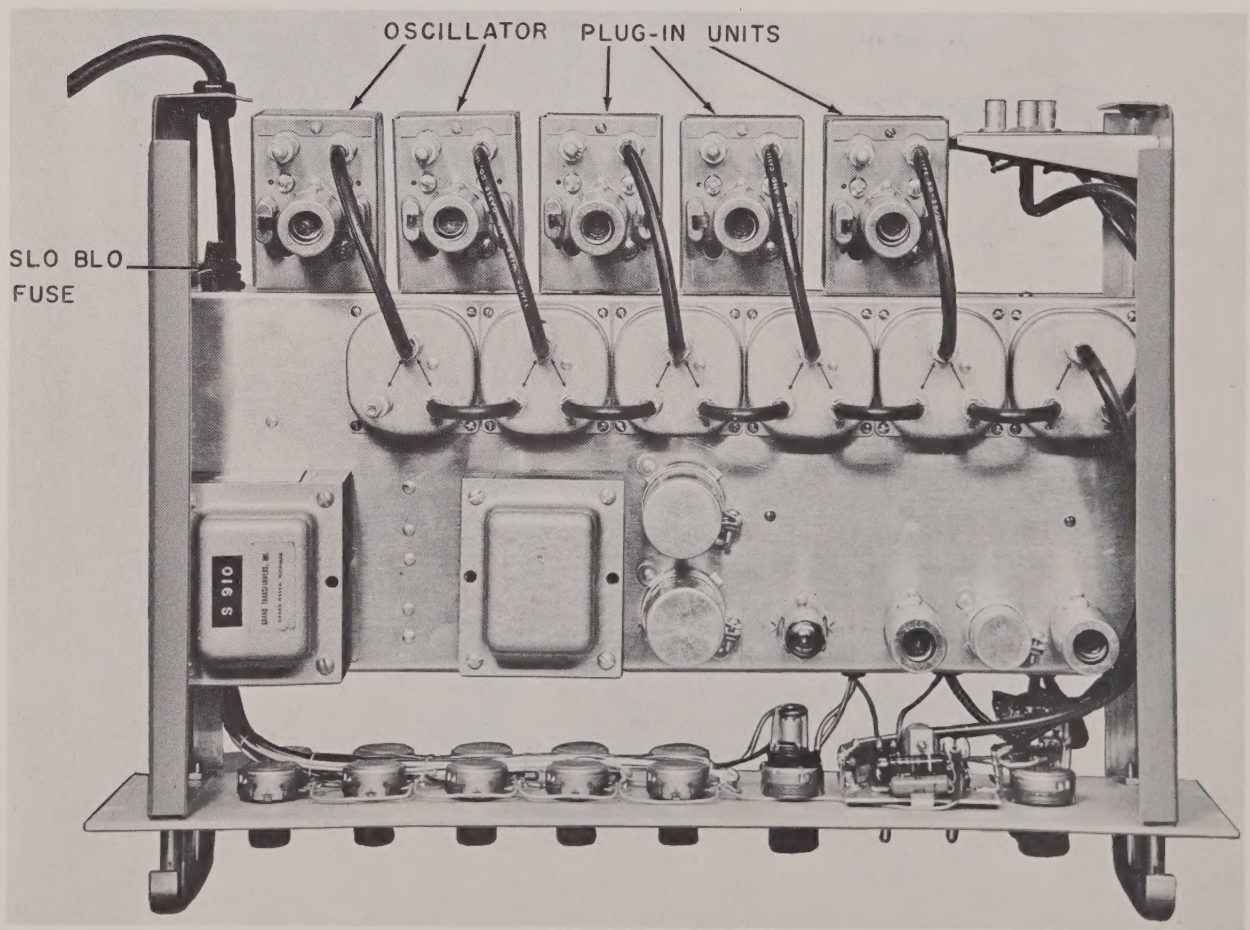



**Figure 2—Jerrold Crystal-Controlled Marker Generator—Model CM-10**
**Top view of chassis showing oscillator plug-in units and location of Slo-Blo Fuse.**

# MARKER GENERATORS Models CM-6, CM-10

## GENERAL DESCRIPTION

The Jerrold Marker Generator Models CM-6 and CM-10 are instruments for superimposing well-defined and precisely crystal-controlled ac or dc frequency marks in the range of 2 mc to 100 mc (Model CM-6) or 100 kc to 100 mc (Model CM-10) upon an oscilloscope display of frequency response curves from electronic devices under test.

Outstanding features:

1. Plug-in oscillator units with individual amplitude control for each oscillator.
2. "Birdie-bypass" circuitry.
3. Display of up to 20 marks.
4. Choice of ac or dc marks.
5. Built-in scope preamplifier.

The portable Model CM-6 comes factory-equipped with three oscillator plug-in units, each unit having two crystal-controlled oscillator sections. The rack-mounted Model CM-10 has five such oscillator plug-in units. (See Figures 1 and 2 on opposite page.)

Additional plug-in units, or single crystals for individual oscillator sections, covering different frequencies, can be ordered and supplied to customers' specifications.

The instruments can thus be programmed to provide any desired fundamental frequencies between 2 mc and 100 mc in the case of Model CM-6, and between 100 kc and 100 mc in the case of Model CM-10.

The oscillator output amplitude can be individually controlled and varied from the front panel.

With the "birdie-bypass" method of marker insertion, the pips are not passed through the equipment under test. The marks therefore remain unaffected by the inherent characteristics of that equipment. Furthermore, marks superimposed on the response curves can have either ac character, extending both above and below the curve, or dc character, extending only above the response curve. In addition, Model CM-10 has a MARKER SHAPE control allowing to vary the bandwidth of the marker pips.

The instruments have a built-in oscilloscope preamplifier and a high-pass filter. This feature assures clear visibility on the scope of low-amplitude sweep voltages, for which otherwise an external scope preamplifier would be required.

The versatility of both Models is further enhanced by the provision on the front panels for connecting two external marker generators. For these too, "birdie-bypass" circuitry is applied.

Models CM-6 and CM-10 are housed in well-ventilated, sturdy cabinets. Model CM-6 is portable and very convenient for field and bench use. Model CM-10 is designed for rack-mounting, with four screws provided for mounting on standard rack constructions. Two handles on the front panel enable easy pullout.

## SPECIFICATIONS

### OVERALL FREQUENCY RANGE

CM-6: 2 mc to 100 mc

CM-10: 100 kc to 100 mc

### OSCILLATOR FREQUENCY RANGES

High-High, Both Models: 2 x 20 mc to 100 mc

Low-High, CM-6: 2 mc to 20 mc + 20 mc to 100 mc

CM-10: 100 kc to 20 mc + 20 mc to 100 mc

Low-Low, CM-6: 2 x 2 mc to 20 mc

CM-10: 2 x 100 kc to 20 mc

### MARKS

Up to 10 crystal-controlled marks;

Up to 20th harmonic;

Wideband marks at 30 kc;

Narrowband marks at 10 kc;

(Model CM-10 allows for variation in marker bandwidth within wide-band and narrow-band ranges)

ac marks for oscilloscopes;

dc marks for X-Y plotters;

Additional inputs for up to two external marker generators

### PREAMPLIFIER

| | |
|---|---|
| GAIN: | 40 db, fixed |
| FLATNESS: | ± 1 db from 10 cycles to 20 kilocycles |
| S/N RATIO: | 1:1, with 20 microvolts input, giving hum and noise output level of less than 3 millivolts |
| MAX. UNDISTORTED OUTPUT: | 5 volts into 100 k ohms, or higher |

## POWER SUPPLY

| | |
|---|---|
| INPUT: | 115/230 volts, 60/50 c.p.s. ac |
| OUTPUT: | Plate voltages, CM-6: 215 v, 150 v<br>CM-10: 260 v, 150 v |
| | Filament voltages,<br>CM-6: 6.3 v ac, 5 v dc<br>CM-10: 6.3 v ac, 12.6 v dc |
| RECTIFICATION: | Silicon Diodes, full-wave Rectifier, for plate voltages |
| | Silicon Diode, half-wave Rectifier, for filament voltages of Marker Amplifier and Preamplier |
| STABILIZATION: | Plate voltage fluctuations compensated for by OA2 stabilizer tube |

## TUBE COMPLEMENT

CM-6: (1) OA2, (2) 12AY7
CM-10: (1) OA2, (2) 12AY7, (1) 6CM6
OSCILLATOR PLUG-IN UNITS: (1) 12AT7, each

## DIMENSIONS:

CM-6: 15″ x 11″ x 7½″
CM-10: 19″ x 14″ x 5″

## NET WEIGHT

CM-6: 21 lbs. CM-10: 24 lbs.

## SHIPPING WEIGHT

CM-6: 31 lbs. CM-10: 34 lbs.

**Figure 3—Jerrold Crystal-Controlled Marker Generator—Model CM-6 Front Panel View**

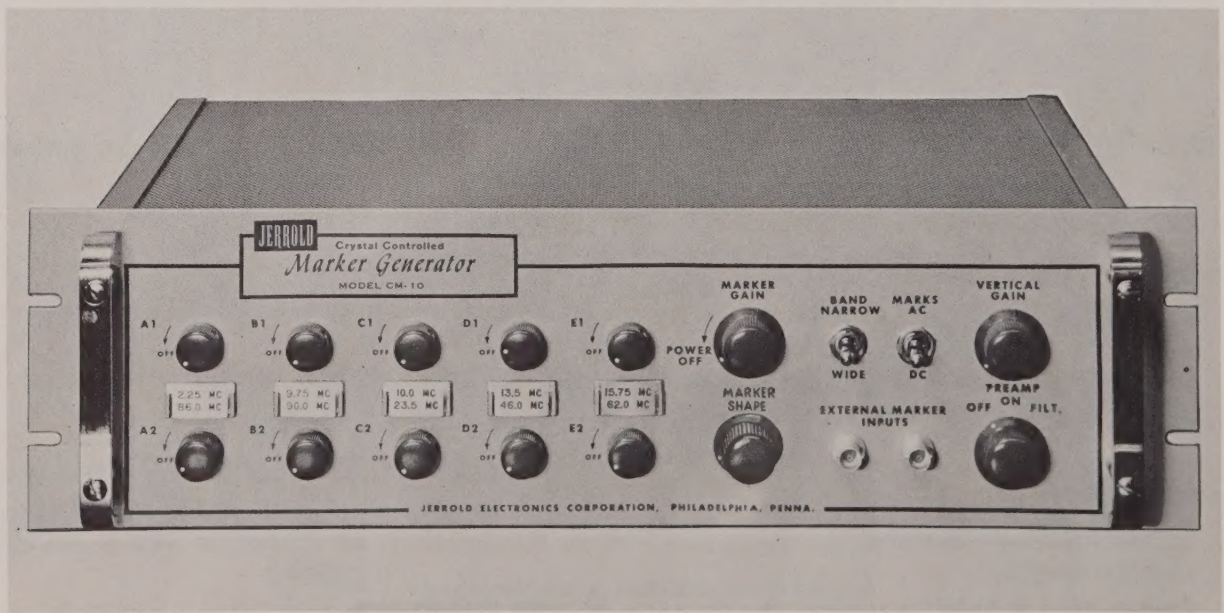



**Figure 4—Jerrold Crystal-Controlled Marker Generator—Model CM-10 Front Panel View**

Figure 6—Jerrold Marker Generators—Models CM-6 and CM-10, Controls Chart

## CHART OF CONTROLS

(Compare Figures 3, 4, 5)

| DESIGNATION | POSITION | FUNCTION |
|---|---|---|
| | OFF | Oscillators cut off |
| **A-1, A-2, B-1, B-2, C-1, C-2, D-1, D-2, E-1, E-2, (CM-10)** | Clockwise turn | Oscillator amplitudes are increased |
| | Counter-clockwise turn | Oscillator amplitudes are decreased |
| **A-1, A-2, B-1, B-2, C-1, C-2 (CM-6)** | | |
| **MARKER GAIN** | POWER OFF | Power switch incorporated in control disconnects set from power source |
| | Clockwise turn | Power on; output levels of all oscillator signals previously selected by A-1, A-2, etc. are increased |
| | Counter-clockwise turn | Output levels of all oscillator signals are decreased |
| **MARKER SHAPE** (CM-10 only) | Clockwise and Counter-clockwise | Increases and decreases bandwidth of markers within range selected by NARROW-WIDEBAND switch |
| **PREAMP** | OFF | Input signal from sweep generator is not amplified |
| | ON | Input signal from sweep generator is amplified |
| | FILT | High-pass filter is inserted into preamplifier network |
| **VERTICAL GAIN** | (This control is continuously functional independent of PREAMP control position) clockwise and counter-clockwise | Varies the vertical gain of the signal output |
| **AC DC MARKS** | AC | Selects markers with ac characteristics |
| | DC | Selects markers with dc characteristics |
| **NARROW-WIDEBAND** | NARROW | Selects markers with 10 kc bandwidth |
| | WIDE-BAND | Selects markers with 30 kc bandwidth |

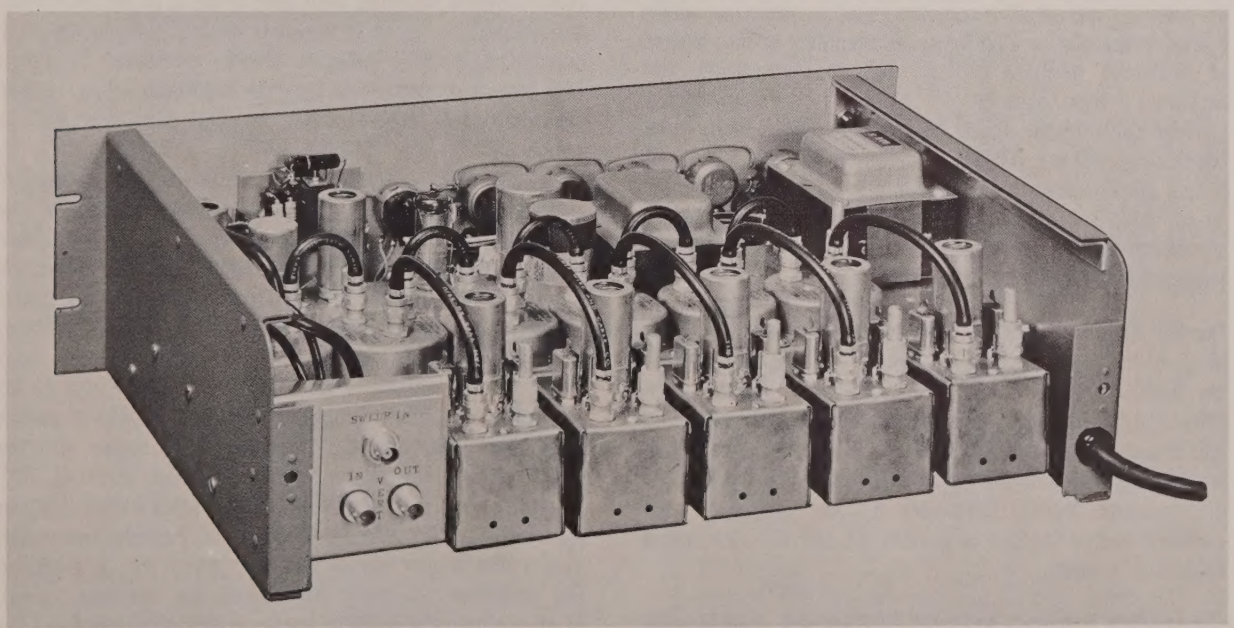


Figure 5—Jerrold Crystal-Controlled Marker Generator—Model CM-10
Rear View of Chassis

# CIRCUIT DESCRIPTION

**(Refer to schematic circuit diagrams, Figures 13 through 17.)**

## 1. POWER SUPPLY

Both models, CM-6 and CM-10, can be energized either from a 115 v 60 c.p.s. or a 230 v, 50 c.p.s. ac source. Silicon diodes provide full-wave rectification for well filtered plate voltages with voltage fluctuations compensated for by a stabilizer tube. In addition, separately rectified filament voltages are supplied for the marker amplifier and preamplier stages to minimize hum. A 750 mA SLO-BLO fuse protects the instruments against surge currents.

NOTE: If the instruments are to be energized from a 230 v ac source, the fuse has to be replaced by a 375 mA type.

## 2. OSCILLATORS

**a. GENERAL.** The shielded plug-in oscillator units have two crystal-controlled oscillator sections each.

Model CM-6, equipped to meet standard requirements, has three plug-in units: two containing two high-frequency oscillator sections each, the third unit containing one high-frequency and one low-frequency oscillator section.

Model CM-10 has 5 plug-in units, each unit having one high-frequency and one low-frequency oscillator section.

**b. OSCILLATOR CIRCUITS.** The high-frequency oscillator circuit covers the range from 20 mc to 100 mc. A 12AT7 twin-triode is employed in a grounded plate "Hartley" circuit. The crystals operate in a series mode from the cathodes to taps on coils L-1 and L-2. The coils are tuned to their respective crystal frequencies or slightly above, according to the number of harmonics desired. Trimmers C-8 and C-9 couple the oscillator energy to connectors J-1 and J-2 and thence via shielded cable to the mixer. All oscillator plug-in units have terminating resistors (TR-72F) mounted on J-2. If it is desired to use one oscillator in one unit as a side-band generator for an oscillator in a second unit, the terminating resistor in the second unit has to be removed and J-1 on the first unit to be jumpered to J-2 on the second unit.

The low frequency oscillator unit covers a range from 2 mc to 20 mc for Model CM-6, and a range from 100 kc to 20 mc for Model CM-10. This circuit is a modified grounded-plate "Pierce" type. Here the crystals operate in parallel mode from grid to cathode. L-1 and L-2 again are tuned to their respective crystal frequencies or slightly above, according to the number of harmonics desired. Capacitors C-4 and C-7 couple the oscillator energy through connectors J-1 and J-2 via shielded cable to the mixer.

The low-high frequency oscillator unit has one low-frequency section and one high-frequency section. The circuit arrangement is a combination of the circuits used in the low-frequency and high-frequency oscillators described above. Here, one half of the twin-triode operates in the modified "Pierce," the other half in the modified "Hartley" circuit.

For Model CM-6 this plug-in unit covers the range of 2 mc to 20 mc and from 20 mc to 100 mc.

For Model CM-10 the ranges are from 100 kc to 20 mc and from 20 mc to 100 mc.

## 3. MIXER STAGE

All mixers, including that for the external marker generators, are identical in design. The oscillator energies enter through the relevant J connectors and are coupled through capacitors to their respective crystal diodes. A sample of the sweep output voltage, entering at SWEEP IN, passes through all mixers. A portion of this output is fed through voltage dividers to the diodes. The dividers also prevent the oscillator voltages in the mixer from affecting the sweep output voltage of the equipment under test, and the mixers from affecting each other through unwanted beat frequencies.

The difference frequencies obtained from the mixer stage are filtered through rc blocks and fed to the marker amplifier stage.

## 4. MARKER AMPLIFIER

In the marker amplifier stage the signals from the mixers are combined on the grid of one section of a 12AY7 twin triode. There they are amplified, with their 10 kc or 30 kc component determined by a choke in the plate circuit. In Model CM-10 a potentiometer (R-51, MARKER SHAPE), connected in series with the choke, permits to vary the bandwidth of the marker pip within the range selected by the NARROW-WIDEBAND switch (SW-2). The MARKER GAIN potentiometer (R-32 in Model CM-6, R-52 in Model CM-10) controls the grid voltage of the second 12AY7 triode, permitting to vary the size of the marker pip. In Model CM-6 the amplified signal from the plate of the second triode is fed via the NARROW-WIDEBAND switch directly to the primary of transformer T-2. In Model CM-10 the signal voltage is further amplified in a 6CB6 pentode, from where it is passed to T-2. This transformer is self-resonant at 30 kc and is tuned to 10 kc by switching an RC block into the circuit through the NARROW-WIDEBAND switch SW-2. The secondary of T-2, connected in series with the signal voltage, injects the markers via the VERTICAL OUT connector upon the trace. With AC-DC MARKS switch (SW-3 in Model CM-6, SW-4 in Model CM-10) in DC position, two diodes rectify the marker signal and a positive pip is produced (its amplitude extending only above the response curve). In AC position, the signal character is left undisturbed and the marker pip amplitude extends above and below the response curve.

### 5. PREAMPLIFIER

The preamplifier is a fixed 40 db gain amplifier with stabilized feedback and dc-coupled interstage. Here too a low-noise 12AY7 twin-triode is used. The amplifier is flat to ± 1 db from 10 cycles to 20 kc. Maximum undistorted output is over 5 volts into 100,000 ohms or higher.

A high-pass filter network switched into the preamplifier circuit with PREAMP switch in FILT position, insures that hf response is not affected when the unit is used with delay line patterns. Low amplitude sweep voltages are clearly visible on the trace and no external oscilloscope preamplifier is necessary.

## OPERATION AND MAINTENANCE

### 1. CONNECTING THE MARKER GENERATOR

The Jerrold Marker Generator Models CM-6 and CM-10 can be used with any conventional sweep frequency generator (Jerrold Models 900-A*, 900-B,* or equivalent) and a laboratory oscilloscope. Fig. 7, shows the general hook-up of the marker generator. Fig. 6, the chart of controls should be carefully studied to become acquainted with the various control functions.

Apart from the controls listed there, three main connections are provided. On Model CM-6 they are mounted on the front panel, on Model CM-10 at the rear of the chassis (compare Fig. 3 and Fig. 5).

1. VERTICAL IN for connection to the vertical output of the sweep frequency generator.
2. VERTICAL OUT for coupling to vertical output of the oscilloscope.
3. SWEEP IN for connection to a T connector (JERROLD MODEL TO-20 or equivalent), the other two legs of which are connected respectively to the rf output of the sweep frequency generator and to the electronic gear under test.

NOTE: Shields of interconnecting cables must be well grounded!

Two further connectors EXTERNAL MARKER INPUTS are provided on the front panels of both Models CM-6 and CM-10, for connecting one or two external marker generators whenever required.

*Request Tech Data Sheets from Jerrold Electronics Corporation, Industrial Products Division, 15th & Lehigh, Phila. 32, Pa. Wire, phone or call.

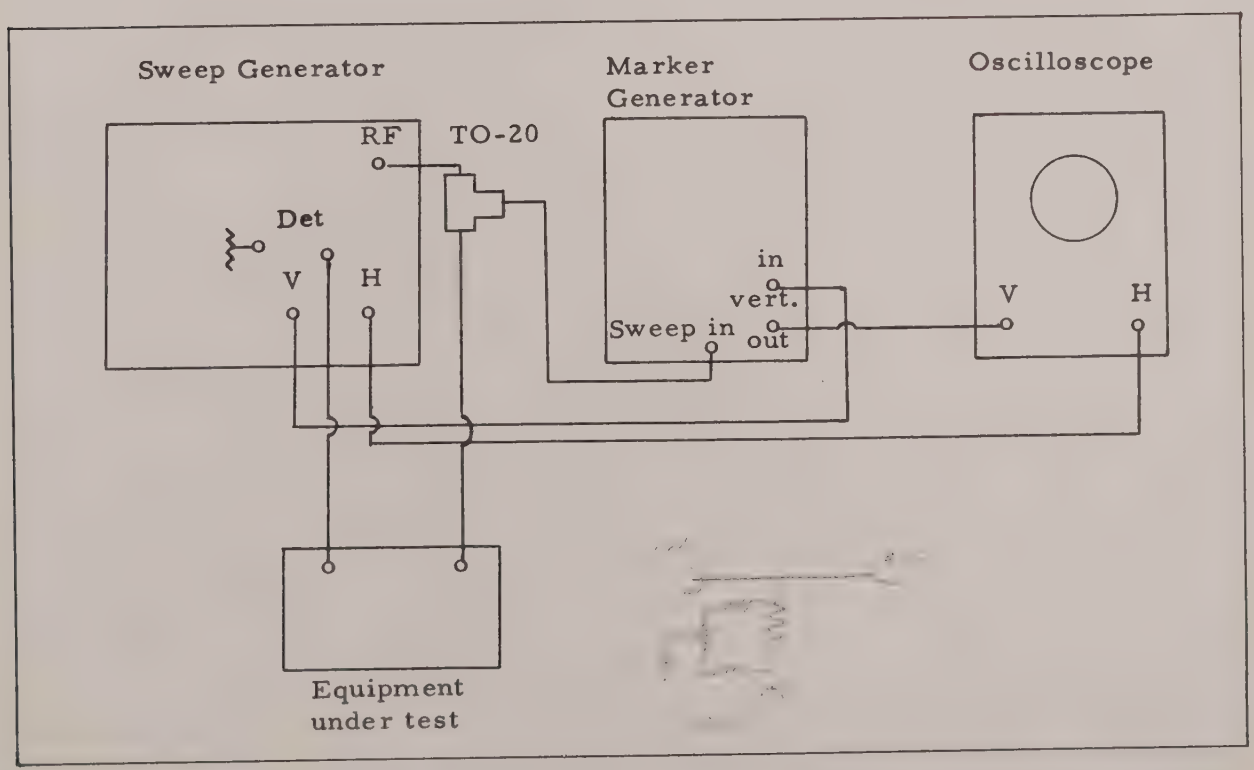



Figure 7—General Hook-Up of Marker Generator

Fig. 8 shows typical scope presentations of frequency response curves having large bandwidth and with well-defined marker pips superimposed at pertinent points across the curves.

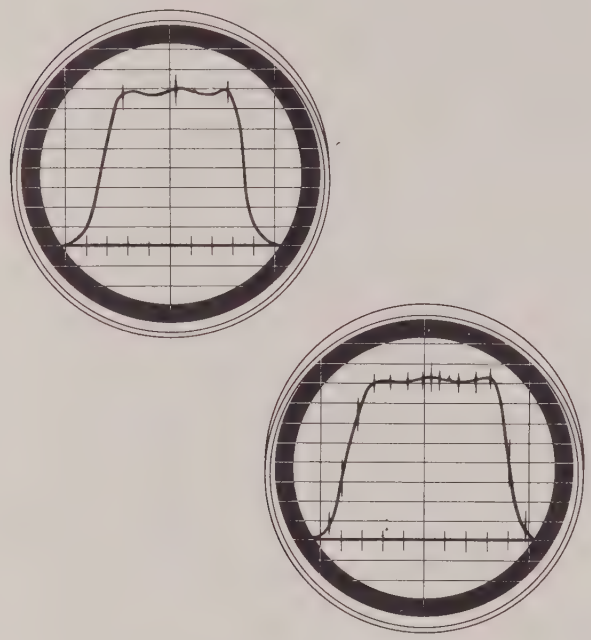


**Figure 8—Typical Response Curves with Marker Pips**

## 2. CHECK-OUT OF THE MARKER GENERATOR

For proper check-out of the instruments, voltage and resistance measurements charts (Figs. 9 thru 12) are given for both Models CM-6 and CM-10. The charts should be used in conjunction with the relevant schematic circuit diagrams. Slight variations in tube characteristics may vary voltage readings to a minor degree.

| TUBE PIN # | MARKER GENERATOR | | | OSCILLATOR |
|---|---|---|---|---|
| | V-1 | V-2 | V-3 | V-1 |
| 1 | 150 V | 225 V | 115 V | 0-185 V* |
| 2 | | 0 | 0 | —0.4 V |
| 3 | | 4 V | 100 V | 0 |
| 4 | | 12 V (DC) | 12 V (DC) | 6.3 V (AC) |
| 5 | 150 V | 0 | 0 | 6.3 V (AC) |
| 6 | | 225 V | 115 V | 0-185 V* |
| 7 | 0 | 0 | 0 ~~115 V~~ | —0.4 V |
| 8 | | 4 V | 115 V | 0 |
| 9 | | | | 0 |

*Voltage measurements vary with operation of individual oscillator amplitude control, all others at zero output position.
NOTE: All measurements were made with a 20,000 ohms per volt instrument and with all controls set at maximum counter-clockwise position.

**Figure 9—Jerrold Crystal-Controlled Marker Generator Model CM-6 Voltage Measurements Chart**

| TUBE PIN # | MARKER GENERATOR | | | OSCILLATOR |
|---|---|---|---|---|
| | V-1 | V-2 | V-3 | V-1 |
| 1 | 20 K* | 15 K* | 125 K* | 10 K |
| 2 | | 95 K | inf. | 100 K |
| 3 | | 1 K | inf. | 0 |
| 4 | | 5 | 5 | 0 |
| 5 | 20 K* | 0 | 0 | 0 |
| 6 | | 15 K* | 50 K* | 10 K |
| 7 | 0 | 2 | 125 K* | 100 K |
| 8 | | 1 K | 100 K* | 0 |
| 9 | | | | 0 |

*Measurements will vary with capacitor charge.
NOTE: All measurements were made with a 20,000 ohms per volt instrument and with all controls set at maximum counter-clockwise position.

**Figure 10—Jerrold Crystal-Controlled Marker Generator Model CM-6 Resistance Measurements Chart**

| TUBE PIN # | MARKER GENERATOR | | | | | | | | OSCILLATOR | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | 0A-2 | | 6CB6 | | 12AY7 | | 12AY7 | | 12AT7 | | | |
| | V-1 | | V-2 | | V-3 | | V-4 | | V-1A (Lo) | | V-1B (Hi) | |
| | A | B | A | B | A | B | A | B | A | B | A | B |
| 1 | 150 dc | 150 dc | 0 | 0 | 115 dc | 110 dc | 115 dc | 60 dc | X | X | 0 | 195 dc |
| 2 | | | 2.0 dc | 1.8 dc | —.14 dc | —.14 dc | 0 | 0 | X | X | —.37 dc | —.37 dc |
| 3 | | | 6.3 ac | 6.3 ac | Gnd. | Gnd. | 105 dc | —.42 dc | X | X | 0 | 0 |
| 4 | | | Gnd. | Gnd. | 12.6 dc | 12.6 dc | 12.6 dc | 12.6 dc | 6.3 ac | 6.3 ac | 6.3 ac | 6.3 ac |
| 5 | 150 dc | 150 dc | 260 dc | 230 dc | Gnd. | Gnd. | Gnd. | Gnd. | 6.3 ac | 6.3 ac | 6.3 ac | 6.3 ac |
| 6 | | | 135 dc | 120 dc | 140 dc | 125 dc | 115 dc | 125 dc | 0 | 210 dc | X | X |
| 7 | Gnd. | Gnd. | Gnd. | Gnd. | 0 | 0 | 115 dc | 60 dc | —.30 dc | —.30 dc | X | X |
| 8 | | | | | 1.4 dc | 1.3 dc | 115 dc | 60 dc | 0 | 0 | X | X |
| 9 | | | | | 6.3 dc | 6.3 dc | 6.3 dc | 6.3 dc | Gnd. | Gnd. | Gnd. | Gnd. |

NOTES: 1. All measurements to ground.
2. All ac measurements made with moving vane dynamometer type instrument.
3. All dc measurements made with 20,000 ohms/volt type instrument.
4. Measurements under A made with all controls at minimum position.
5. Measurements under B made with all controls at maximum position.

**Figure 11—Jerrold Crystal-Controlled Marker Generator Model CM-10, Voltage Measurements Chart.**

| TUBE PIN # | MARKER GENERATOR | | | | OSCILLATOR | |
|---|---|---|---|---|---|---|
| | 0A2 | 6CB6 | 12AY7 | 12AY7 | 12AT7 | |
| | V-1 | V-2 | V-3 | V-4 | V-1A | V-1B |
| 1 | Charge 20 K to 40 K | 100 K | 40 K | 120 K | X | 20 K |
| 2 | | 220 Ω | 100 K | 3 M | X | 100 K |
| 3 | | 0.1 Ω | 0 | 560 Ω | X | 0.4 Ω |
| 4 | | 0 | 6 Ω | 6 Ω | 0.1 Ω | 0.1 Ω |
| 5 | Charge 20 K to 40 K | 22 K | 0 | 0 | 0.1 Ω | 0.1 Ω |
| 6 | | 80 K | 60 K | 60 K | 20 K | X |
| 7 | 0 | 0 | 100 K | 120 K | 100 K | X |
| 8 | | | 560 K | 85 K | 0.1 | X |
| 9 | | . | 4.5 | 4.5 | 0 | 0 |

NOTES: 1. All measurements to ground.
2. All controls at maximum position.
3. Oscillator unit plugged in.
4. Switch PREAMP in ON position.

**Figure 12—Jerrold Crystal-Controlled Marker Generator Model CM-10, Resistance Measurements Chart**

### 3. FIELD MAINTENANCE

Under normal usage and with proper application, Jerrold Marker Generator Models CM-6 and CM-10 give accurate, trouble-free, long-life service.

Replacement parts lists for both models are included in this manual. When ordering any component, the model number, serial number, and the Jerrold part number should be quoted. When replacing a tube, only a new exactly equivalent type of a recognized brand should be substituted.

When replacing rectifier diodes or crystals, proper polarity must be observed.

In case the instrument does not function properly Jerrold's Instrument Service Department should be notified immediately. Give the set's serial number and describe accurately the difficulties encountered. Information will promptly be forwarded, how to correct the instrument, or authorization will be given to return the instrument to the factory.

In case the set is being returned to the factory, it must bear its identification tag. Pack carefully and ship prepaid via express.

See our warranty on page 8.

### 4. ALIGNMENT

**a. GENERAL.** As explained in the circuit description, the low frequency oscillators and the high frequency oscillators have different basic circuitry.

Alignment procedure will therefore differ too; however, any oscillator section may be adjusted for maximum performance and can be aligned to have either rich harmonic content or low harmonic content.

The instruments required for alignment are a high quality sweep generator (such as Jerrold Models 900-A or 900-B) and a high quality oscilloscope. In the alignment procedure given here, the set-up presupposes the use of a Jerrold Model 900-A Sweep Generator. When using a different instrument, the control functions will be the same, though the control designations may differ for different models.

**b. EQUIPMENT SET-UP**

(1) Set the sweep generator to VHF range of 0-400 mc; blanking ON; automatic gain control ON; set marker generator MARKER GAIN control at maximum counter-clockwise position.

(2) Set oscilloscope vertical sensitivity to 30 mV/cm.

(3) Set all oscillator amplitude controls (A-1, A-2, B-1, B-2, etc.) on marker generator at OFF (maximum counter-clockwise).

(4) Set NARROW-WIDEBAND switch on marker generator at WIDEBAND.

(5) Set AC-DC MARKS switch on marker generator at AC.

(6) Set PREAMP control on marker generator at OFF.

(7) Set VERTICAL GAIN control on marker generator so that a 4 cm deflection appears on the scope.

(8) Now turn MARKER GAIN control on marker generator just below noise level.

NOTE: Allow at least ½ hour for proper warm-up of equipment before proceeding with alignment.

**c. ALIGNMENT PROCEDURE**

(1) Low-frequency oscillator with many harmonics:

(a) Turn L-1 to maximum counter-clockwise position, C-6 to mid-range.

(b) Turn L-1 clockwise until maximum number of harmonics appear.

(c) Turn oscillator amplitude control from OFF position slowly in clockwise direction until oscillations start. Highest pip to show on the scope should indicate fundamental frequency. If second harmonic is highest, L-1 is turned in too far; adjust as necessary.

(2) Low-frequency oscillator with few harmonics:

(a) Turn C-6 to minimum coupling position.

(b) Turn L-1 clockwise until oscillations cease, then counter-clockwise until oscillations start. This will give point with minimum number of harmonics.

(3) High-frequency oscillator with many harmonics:

(a) Turn L-2 and L-4 to maximum clockwise position with C-9 at about midrange.

(b) Turn oscillator amplitude control to maximum clockwise position.

(c) Turn L-2 slowly clockwise until oscillator drops out, then back up slowly until oscillator drops in.

(d) Turn L-4 clockwise until circuit goes into self-oscillation; this becomes apparent by pip walking while L-2 is being tuned.

(e) Back off L-4 until L-2 can be tuned through entire range without the pip walking.

(f) Switch sweep generator to UHF and turn VERTICAL GAIN control on marker generator to display pattern on CRT screen.

(g) Turn L-2 counter-clockwise and vary C-9 until pips are maximum size.

(h) Return sweep generator to VHF, turn oscillator amplitude control to OFF.

(i) Turn oscillator amplitude control slowly clockwise until oscillations set in. (Oscillations should start with about 25% rotation of control.) Largest pip should indicate fundamental frequency.

(4) High-frequency oscillators with few harmonics:

(a) Set up oscillator as in (3) above.

(b) Reduce coupling of C-9 to minimum.

(c) Set oscillator amplitude control at approx. 50% rotation and turn L-4 counter-clockwise until oscillations cease; then return until oscillations start again.

(d) Turn L-2 clockwise until oscillations cease, then counter-clockwise until oscillations start again. This is point of minimum number of harmonics.

# Warranty

1. The Manufacturer warrants this test equipment, including all its component parts (except tubes, semi-conductors and relays), to be free from defects in material and workmanship under normal use and service, its obligations being limited to making good at its factory any part or parts so warranted which shall within one year after delivery by the Manufacturer be returned to it with transportation charges prepaid and which its examination shall disclose to its satisfaction to have been thus defective.

2. The benefit of this warranty shall not apply to any instrument or part thereof which shall have been repaired or altered outside of a repair unit, authorized or approved by the Manufacturer, in any way so as in the judgment of the Manufacturer to affect its stability and reliability, nor which has been subject to misuse, neglect, accident or improper application.

3. The Manufacturer disclaims any warranty other than as specifically set forth herein, and may discontinue models or alter their specifications without any obligation to incorporate any modifications in equipment previously sold.

## NOTES

# JERROLD MARKER GENERATOR MODEL CM-6
# REPLACEMENT PARTS LIST

| Item | JERROLD Part # | QTY | Schematic Designation | Mfr's Name | Mfr's # | Description | | |
|---|---|---|---|---|---|---|---|---|
| | | | | | | **CAPACITORS** | | |
| 1 | 121-014 | 4 | C-3, 8, 13, 18 | Centralab | TCZ-20 | 20 mmf, 600 V, ± 2%, temp. comp. | | |
| 2 | 122-006 | 4 | C-4, 9, 14, 19 | Quality Comp. | QC-series | 1.5 mmf, ± 10%, gimmick | | |
| 3 | 122-010 | 4 | C-5, 10, 15, 20 | Quality Comp. | QC-series | 3.3 mmf, ±10%, gimmick | | |
| 4 | 122-047 | 4 | C-6, 11, 16, 2 | Centralab | DD-560 | 56 mmf, 1000 V, ±10%, cer. disc | | |
| 5 | 124-031 | 4 | C-7, 12, 17, 22 | Erie | BPD Type | 0.01 mfd, 600 V, cer. disc | | |
| 6 | 125-030 | 2 | C-23, 27 | Sprague | 4TM-S15 | 0.015 mfd, 400 V | | |
| 7 | 125-029 | 1 | C-24 | Sprague | 4Tm-D2 | 0.002 mfd, 400 V, paper moulded | | |
| 8 | 125-016 | 1 | C-25 | Sprague | 4TM-D1 | 0.001 mfd, 400 V, paper moulded | | |
| 9 | 125-006 | 1 | C-26 | Sprague | 6Tm-T4 | 0.004 mfd, 600 V, paper moulded | | |
| 10 | 124-101 | 1 | C-28 | Erie | 838-500 | 100 mmf, 150 V, ± 20%, cer. disc | | |
| 11 | 123-113 | 2 | C-29, 35 | Erie | GP-2-3-31 | 470 mmf, 500 V, ± 10% | | |
| 12 | 125-031 | 1 | C-30 | Aerovox | P-123 ZB | 0.047 mfd, 200 V, paper moulded | | |
| 13 | 125-013 | 1 | C-31 | Cornell-Dubilier | CUB-451 | 0.01 mfd, 400 V, paper moulded | | |
| 14 | 125-032 | 1 | C-32 | Sprague | 2TM-P15 | 0.15 mfd, 200 V, paper moulded | | |
| 15 | 127-700 | 1 | C-33 | Pyramid | TM series | 20 + 10 mfd, 300-350 V, electrolytic | | |
| 16 | 125-038 | 1 | C-34 | Cornell-Dubilier | PM4P25 | 0.25 mfd, 400 V, paper moulded | | |
| 17 | 127-506 | 1 | C-36 | Mallory | WP-068 | 1500 mfd, 50 V, electrolytic | | |
| 18 | 127-601 | 1 | C-37A, B | Pyramid | TM-D-40 | 40 + 40 mfd, 350 V, electrolytic | | |
| | | | | | | **CHOKE** | | |
| 19 | 823-015 | 1 | CH-1 | Jerrold | 823-015 | Special Assembly | | |
| | | | | | | **CRYSTAL DIODES** | | |
| 20 | 139-113 | 4 | CR-1, 2, 3, 4 | Kemtron | K3A | Silicon diode | | |
| | | | | | | **FUSE** | | |
| 21 | 101-340 | 1 | FS-1 | SLO-BLO | 3AG | 750 mA | | |
| | | | | | | **PILOT LIGHT** | | |
| 22 | 102-002 | 1 | PL-1 | GE | #47 | 6, 3 V, 300 mA | | |
| | | | | | | **RECTIFIER DIODES** | | |
| 23 | 137-706 | 4 | SiR-1, 2, 3, 4 | Sarkes-Tarzian | M-150 | 150 mA, silicon | | |
| 24 | 137-707 | 1 | SiR-5 | Sarkes-Tarzian | M-500 | 500 mA, silicon | | |

| | | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| | | | | | | **RESISTORS** | | |
| 25 | 112-112 | 1 | R-1 | Allen-Bradley | "HB" series | 10 Ω, 2 W, ± 10% | | |
| 26 | 113-008 | 1 | R-2 | True-Ohm | "FRL" series | 1750 Ω, 10 W, wire wound | | |
| 27 | 112-448 | 1 | R-3 | Allen-Bradley | "HB" series | 4700 Ω, 2 W, ± 10% | | |
| 28 | 118-062 | 6 | R-4 thru 9 | Allen-Bradley | "J" series | Pot. 100 K, 2 W, ⅜" shaft & bush | | |
| 29 | 112-119 | 1 | R-10 | Allen-Bradley | "EB" series | 10 Ω, ½ W, ± 10% | | |
| 30 | 112-173 | 3 | R-11, 12, 13 | Allen-Bradley | "EB" series | 33Ω, ½ W, ± 10% | | |
| 31 | 112-698 | 4 | R-14, 18, 22, 26 | Allen-Bradley | "EB" series | 0.47 M, ½ W, ± 10% | | |
| 32 | 112-488 | 4 | R-15, 19, 23, 27 | Allen-Bradley | "EB" series | 10 K, ½ W, ± 10% | | |
| 33 | 112-446 | 4 | R-16, 20, 24, 28 | Allen-Bradley | "EB" series | 4700 Ω, ½ W, ± 10% | | |
| 34 | 112-362 | 7 | R-17, 21, 25, | | | | | |
| | | | 29, 31, 33, 44 | Allen-Bradley | "EB" series | 1 K, ½ W, ± 10% | | |
| 35 | 112-614 | 3 | R-30, 43, 47 | Allen-Bradley | "EB" series | 100 K, ½ W, ± 10% | | |
| 36 | 118-064 | 1 | R-32 | Allen-Bradley | "JS" series | Pot. 100 K, 2 W, with switch, "A" taper, ⅜" shaft & bush | | |
| 37 | 112-530 | 1 | R-34 | Allen-Bradley | "EB" series | 22 K, ½ W, ± 10% | | |
| 38 | 112-572 | 2 | R-35, 36 | Allen-Bradley | "EB" series | 47 K, ½ W, ± 10% | | |
| 39 | 118-063 | 1 | R-37 | Allen-Bradley | "J" series | Pot. 100 K, 2W, "A" taper, ⅜" shaft & bush | | |
| 40 | 112-542 | 1 | R-38 | Allen-Bradley | "EB" series | 27 K, ½ W, ± 10% | | |
| 41 | 112-656 | 1 | R-39 | Allen-Bradley | "EB" series | 220 K, ½ W, ± 10% | | |
| 42 | 112-803 | 1 | R-40 | Allen-Bradley | "EB" series | 3.3 M, ½ W, ± 10% | | |
| 43 | 112-332 | 1 | R-41 | Allen-Bradley | "EB" series | 560 Ω, ½ W, ± 10% | | |
| 44 | 112-590 | 1 | R-42 | Allen-Bradley | "EB" series | 68 K, ½ W, ± 5% | | |
| 45 | 112-527 | 1 | R-45 | Allen-Bradley | "EB" series | 22 K, ½ W, ± 5% | | |
| 46 | 112-563 | 1 | R-46 | Allen-Bradley | "EB" series | 39 K, ½ W, ± 10% | | |
| | | | | | | **SWITCHES** | | |
| 47 | Part of 118-064 | 1 | SW-1 | Allen-Bradley | | Part of Pot. R-32 | | |
| 48 | 162-008 | 2 | SW-2, 4 | Carling | 316-63 | Toggle Switch, DPDT | | |
| 49 | 161-117 | 1 | SW-3 | Jerrold | 161-117 | Wafer Switch | | |
| | | | | | | **TERMINATING RESISTOR** | | |
| 50 | TR-72F | 1 | TR-72F | Jerrold | TR-72F | 75 Ω | | |
| | | | | | | **TRANSFORMERS** | | |
| 51 | 141-104 | 1 | T-1 | Jerrold | 141-104 | Power Transformer | | |
| 52 | 823-013 | 1 | T-2 | Jerrold | 823-018 | Special Assembly | | |
| | | | | | | **TUBES** | | |
| 53 | 132-100 | 1 | V-1 | RCA | OA2 | Voltage Stabilizer | | |
| 54 | 131-402 | 2 | V-2, 3 | RCA | 12AY7 | Twin Triode | | |

# JERROLD MARKER GENERATOR MODEL CM-10
## REPLACEMENT PARTS LIST

| Item | JERROLD Part # | QTY | Schematic Designation | Mfr's Name | Mfr's # | Description | | |
|---|---|---|---|---|---|---|---|---|
| | | | | | | **CAPACITORS** | | |
| 1 | 121-014 | 6 | C-1, 6, 11, 16, 21, 26 | Centralab | TCZ-20 | 20 mmf, 600 V, ± 2% | | |
| 2 | 122-006 | 6 | C-2, 7, 12, 18, 23, 28 | Quality Comp. | QC series | 1.5 mmf, gimmick, ± 10% | | |
| 3 | 122-010 | 6 | C-3, 8, 13, 17, 22, 27 | Quality Comp. | QC series | 3.5 mmf, gimmick, ± 10% | | |
| 4 | 124-047 | 6 | C-4, 9, 14, 19, 24, 29 | Centralab | DD-560 | 56 mmf, 1000 V, ± 10% | | |
| 5 | 124-031 | 6 | C-5, 10, 15, 20, 25, 30 | Erie | BPD type | 0.01 mfd, 600 V, ± 10% | | |
| 6 | 124-049 | 1 | C-31 | Aerovox | D-1 series | 100 mmf, 1000 V, ± 10% | | |
| 7 | 125-001 | 1 | C-32 | Aerovox | 88W | 0.047 mfd, 400 V, ± 20% | | |
| 8 | 125-032 | 1 | C-34 | Sprague | 2TM-P15 | 0.15 mfd, 200 V, ± 10% | | |
| 9 | 127-700 | 1 | C-35 | Pyramid | TM series | 20 + 10 mfd, 300-350 V, electrolytic | | |
| 10 | 125-038 | 1 | C-36 | Cornell-Dubilier | PM-4P25 | 0.25 mfd, 400 V, ± 10% | | |
| 11 | 124-050 | 2 | C-37, 38 | Aerovox | D-1 series | 2000 mmf, 1000 V, ± 10% | | |
| 12 | 126-008 | 1 | C-39 | Arco | CM-15E-151J | 150 mmf, 500 V, ± 5% | | |
| 13 | 126-016 | 1 | C-40 | Arco | CM-19B-751J | 750 mmf, 500 V, ± 5% | | |
| 14 | 126-045 | 1 | C-41 | Arco | CM-19B-821J | 820 mmf, 500 V, ± 5% | | |
| 15 | 125-029 | 1 | C-42 | Sprague | 6TM-D2 | 2000 mmf, 600 V, ± 10% | | |
| 16 | 125-006 | 1 | C-43 | Sprague | 6TM-T4 | 4000 mmf, 600 V | | |
| 17 | | 1 | C-44 | | | special, varies with unit | | |
| 18 | 125-031 | 1 | C-45 | Aerovox | P123ZB | 0.047 mfd, 200 V, ± 20% | | |
| 19 | 124-034 | 1 | C-46 | Erie | 817 BPD | 0.02 mfd, 600 V | | |
| 20 | 125-018 | 1 | C-47 | Sprague | 4TM-S3 | 0.03 mfd, 400 V | | |
| 21 | 127-605 | 1 | C-48A, B | Pyramid | TM-D-40 | 40 + 40 mfd, 350 V, electrolytic | | |
| 22 | 127-506 | 1 | C-49 | Mallory | WP-068 | 1500 mfd, 50 V, electrolytic | | |
| | | | | | | **CHOKES** | | |
| 23 | 143-115 | 2 | Ch-1, 2 | Thorardson | WC-30 | variable (tuned to 75 mH) | | |
| 24 | 143-102 | 1 | CH-3 | Jerrold | 143-102 | broad band ampl. | | |
| | | | | | | **CRYSTAL DIODES** | | |
| 25 | 139-113 | 6 | CR-1 thru 6 | Kemtron | K3A | Silicon Diodes | | |
| 26 | 139-117 | 2 | CR-7, 8 | Kemtron | 1N54 | Ge Diodes | | |
| | | | | | | **FUSE** | | |
| 27 | 101-340 | 1 | FS-1 | SLO-BLO | 3 AG | 750 mA | | |
| | | | | | | **PILOT LIGHT** | | |
| 28 | 102-001 | 1 | PL-1 | GE | #51 | 6.3 V, 0.3 A | | |
| | | | | | | **RECTIFIER DIODES** | | |
| 29 | 137-706 | 4 | SiR-1, 2, 3, 4 | Sarkes-Tarzian | M-150 | 150 mA, silicon | | |
| 30 | 137-707 | 1 | SiR-5 | Sarkes-Tarzian | M-500 | 500 mA, silicon | | |

| | | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| | | | | | | **RESISTORS** | | |
| 31 | 118-062 | 10 | R-1 thru R-10 | Allen-Bradley | "J" series | Pot. 100 K, 2 W, linear, 3/8" shaft & bush | | |
| 32 | 112-173 | 3 | R-11, 12, 13 | Allen-Bradley | "EB" series | 33 Ω, 1/2 W, ± 10% | | |
| 33 | 112-698 | 6 | R-14, 18, 22 | Allen-Bradley | "EB" series | 470 Ω, 1/2 W, ± 10% | | |
| 34 | 112-446 | 6 | R-15, 19, 23, 27, 31, 35 | Allen-Bradley | "EB" series | 4700 Ω, 1/2 W, ± 10% | | |
| 35 | 112-488 | 7 | R-16, 20, 24 | Allen-Bradley | "EB" series | 10 K, 1/2 W, ± 10% | | |
| 36 | 112-362 | 8 | R-17, 21, 25, 29, 33, 37, 45, 50 | Allen-Bradley | "EB" series | 1 K, 1/2 W, ± 10% | | |
| 37 | 112-542 | 1 | R-38 | Allen-Bradley | "EB" series | 27 K, 1/2 W, ± 10% | | |
| 38 | 112-656 | 1 | R-39 | Allen-Bradley | "EB" series | 220 K, 1/2 W, ± 10% | | |
| 39 | 112-803 | 1 | R-40 | Allen-Bradley | "EB" series | 3.3 M, 1/2 W, ± 10% | | |
| 40 | 112-332 | 2 | R-41, 56 | Allen-Bradley | "EB" series | 560 Ω, 1/2 W, ± 10% | | |
| 41 | 112-503 | 1 | R-42 | Allen-Bradley | "EB" series | 13 K, 1/2 W, ± 5% | | |
| 42 | 112-614 | 3 | R-43, 46, 57 | Allen-Bradley | "EB" series | Pot. 100 K, 2 W, "A" taper, 1/2" shaft, 3/8" bush | | |
| 43 | 112-599 | 1 | R-44 | Allen-Bradley | "EB" series | 75 K, 1/2 W, ± 5% | | |
| 44 | 112-536 | 1 | R-47 | Allen-Bradley | "EB" series | 39 K, 1/2 W, ± 10% | | |
| 45 | 118-064 | 1 | R-48 | Allen-Bradley | "J" series | Pot. 100 K, 2 W, with switch | | |
| 45 | 118-063 | 1 | R-49 | Allen-Bradley | "EB" series | 33 K, 1/2 W, ± 10% | | |
| 47 | 118-016 | 1 | R-51 | Allen-Bradley | | Pot. 50 K, 2 W, 3/8" shaft, 1/4" thd. | | |
| 48 | 118-064 | 1 | R-52 | Allen-Bradley | "JS" series | Pot. 100 K, 2 W, "A" taper, 1/2" shaft, 3/8" bush, with switch | | |
| 49 | 112-572 | 1 | R-53 | Allen-Bradley | "EB" series | 47 K, 1/2 W, ± 10% | | |
| 50 | 112-593 | 1 | R-55 | Allen-Bradley | "EB" series | 58 K, 1/2 W, ± 10% | | |
| 51 | 112-278 | 1 | R-58 | Allen-Bradley | "EB" series | 220 Ω, 1/2 W, ± 10% | | |
| 52 | 113-011 | 1 | R-59 | True-Ohm | "ERL" series | 10 K, 10 W, wire wound | | |
| 53 | 112-112 | 1 | R-60 | Allen-Bradley | "HB" series | 10 Ω, 2 W, ± 10% | | |
| | | | | | | **SWITCHES** | | |
| 54 | 161-117 | 1 | SW-1 | Jerrold | 161-117 | Wafer Switch | | |
| 55 | 161-007 | 1 | SW-2 | Carling | 112-63 | Toggle Switch, SPDT | | |
| 56 | 161-008 | 1 | SW-3 | Carling | 316-63 | Toggle Switch, DPDT | | |
| 57 | Part of 118-064 | 1 | SW-4 | | | Part of Pot. R-48 | | |
| | | | | | | **TERMINATING RESISTOR** | | |
| 58 | TR-72F | 1 | TR-72F | Jerrold | TR-72F | 75 Ω | | |
| | | | | | | **TRANSFORMERS** | | |
| 59 | 141-154 | 1 | T-1 | Jerrold | 141-154 | Power Transformer | | |
| 60 | 823-021 | 1 | T-2 | Jerrold | 823-021 | Special Assembly | | |
| | | | | | | **TUBES** | | |
| 61 | 132-100 | 1 | V-1 | RCA | OA2 | Voltage Stabilizer | | |
| 62 | 131-313 | 1 | V-2 | RCA | 6CB6 | Pentode | | |
| 63 | 131-402 | 2 | V-3, V-4 | RCA | 12AY7 | Twin-Triode | | |

# JERROLD OSCILLATOR PLUG-IN UNITS FOR MARKER GENERATOR MODELS CM-6 AND CM-10, REPLACEMENT PARTS LIST

| Item | JERROLD Part # | QTY | Schematic Designation | Mfr's Name | Mfr's # | Description | | |
|---|---|---|---|---|---|---|---|---|
| | | | | | **1. Hi & Hi FREQ. OSCILLATOR** | | | |
| | | | | | **CAPACITORS** | | | |
| 1 | 123-115 | 5 | C-1, 2, 3, 4, 5 | Erie | GP2-332 | 1000 mmf, 600 V | | |
| 2 | | 1 | C-6 | | | special, according to frequency scale | | |
| 3 | | 1 | C-7 | | | special, according to frequency scale | | |
| 4 | 128-505 | 2 | C-8, 9 | Erie | 535 | 0.7-3 mmf, 350 V, trimmer | | |
| | | | | | **CRYSTAL DIODES** | | | |
| 5 | 139-109 | 2 | CR-1, 2 | McCoy | special | special, according to frequency requirement | | |
| | | | | | **RESISTORS** | | | |
| 6 | 112-488 | 2 | R-1, 3 | Allen-Bradley | "EB" series | 10 K, ½ W, ± 10% | | |
| 7 | 112-614 | 2 | R-2, 4 | Allen-Bradley | "EB" series | 100 K, ½ W, ± 10% | | |
| | | | | | **TERMINATING RESISTOR** | | | |
| 8 | TR-72F | 1 | TR-72F | Jerrold | TR-72F | 75 Ω | | |
| | | | | | **TUBE** | | | |
| 9 | 131-400 | 1 | V-1A, B | RCA | 12AT7 | Twin-Triode | | |
| | | | | | **2. Hi & Lo FREQ. OSCILLATOR** | | | |
| | | | | | **CAPACITORS** | | | |
| 1 | 123-115 | 2 | C-1, 3 | Erie | GP2-332 | 1000 mmf, 600 V, ceramic disc | | |
| 2 | | 1 | C-2 | | | special, according to frequency scale | | |
| 3 | 128-505 | 1 | C-4 | Erie | 535 | 0.7-3 mmf, 350 V, trimmer | | |
| 4 | 124-031 | 3 | C-5, 7, 9 | Erie | BPD Type | 0.01 mmf, 600 V, ceramic disc | | |
| 5 | 128-500 | 1 | C-6 | Erie | 557-NPO | 1.5-7 mmf, 350 V, trimmer | | |
| 6 | 126-013 | 1 | C-8 | Arco | CM-15 | 100 mmf, 500 V, ± 5%, mica | | |

| | | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| | | | | | **CRYSTAL DIODES** | | | |
| 7 | 139-109 | 2 | CR-1, 2 | McCoy | special | special, according to frequency requirement | | |
| | | | | | **RESISTORS** | | | |
| 8 | 112-488 | 2 | R-1, 3 | Allen-Bradley | "EB" series | 10 K, ½ W, ± 10% | | |
| 9 | 112-614 | 2 | R-2, 4 | Allen-Bradley | "EB" series | 100 K, ½ W, ± 10% | | |
| | | | | | **TERMINATING RESISTOR** | | | |
| 10 | TR-72F | 1 | TR-72F | Jerrold | TR-72F | 75 Ω | | |
| | | | | | **TUBE** | | | |
| 11 | 131-400 | 1 | V-1A, B | RCA | 12AT7 | Twin-Triode | | |
| | | | | | **3. Lo & Lo FREQ. OSCILLATOR** | | | |
| | | | | | **CAPACITORS** | | | |
| 1 | 126-013 | 2 | C-1, 9 | Arco | CM-15 | 100 mmf, 500 V, ± 5%, mica | | |
| 2 | 124-031 | 5 | C-2, 3, 5, 6, 8 | Erie | BPD Type | 0.01 mfd, 600 V, ceramic disc | | |
| 3 | 128-500 | 2 | C-4, 7 | Erie | 557-NPO | 1.5-7 mmf, 350 V, trimmer | | |
| | | | | | **CRYSTAL DIODES** | | | |
| 4 | 139-109 | 2 | CR-1, 2 | McCoy | special | special, according to frequency requirement | | |
| | | | | | **RESISTORS** | | | |
| 5 | 112-614 | 2 | R-1, 4 | Allen-Bradley | "EB" series | 100 K, ½ W, ± 10% | | |
| 6 | 112-488 | 2 | R-2, 3 | Allen-Bradley | "EB" series | 10 K, ½ W, ± 10% | | |
| | | | | | **TERMINATING RESISTOR** | | | |
| 7 | TR-72F | 1 | TR-72F | Jerrold | TR-72F | 75 Ω | | |
| | | | | | **TUBE** | | | |
| 8 | 131-400 | 1 | V-1A, B | RCA | 12AT7 | Twin-Triode | | |

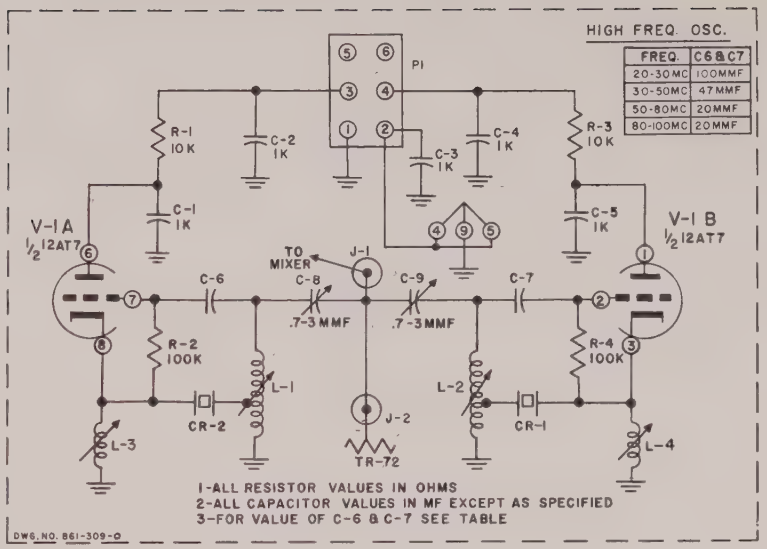



Figure 13

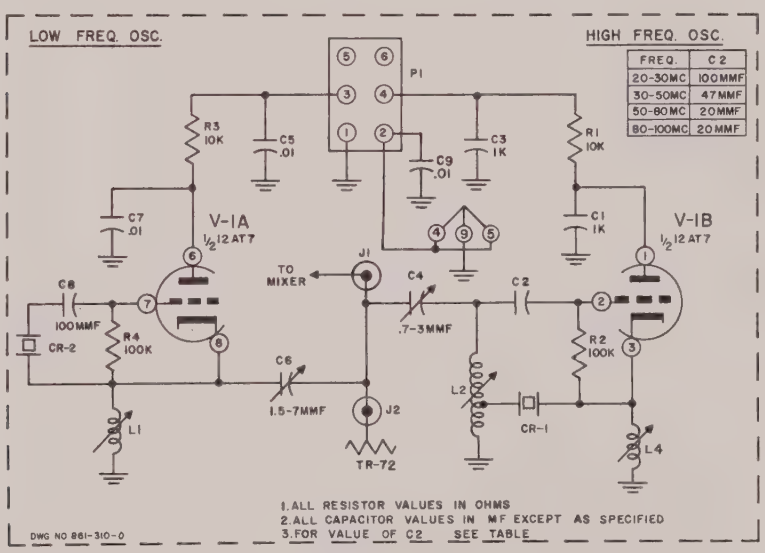



Figure 14

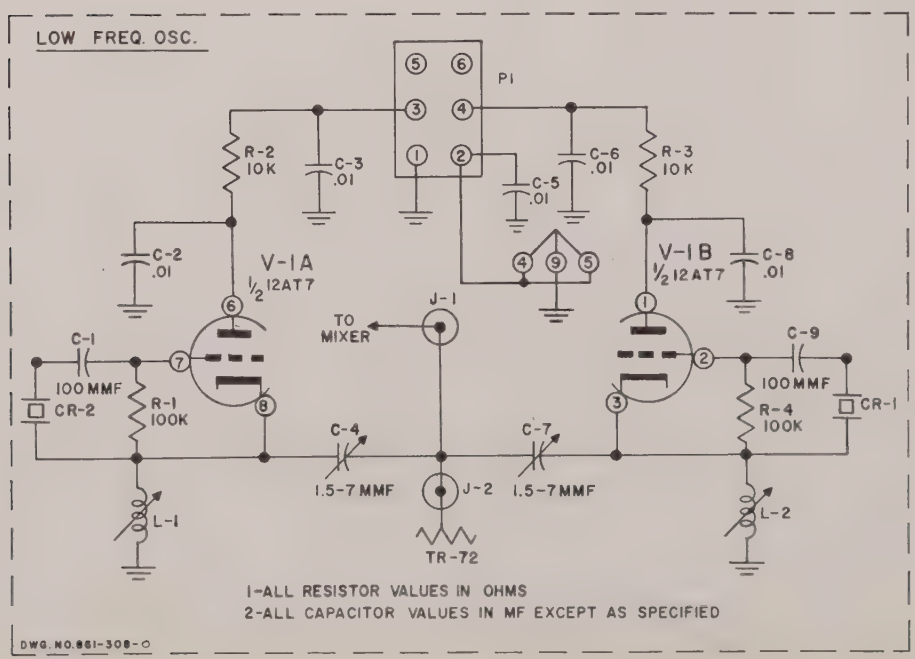
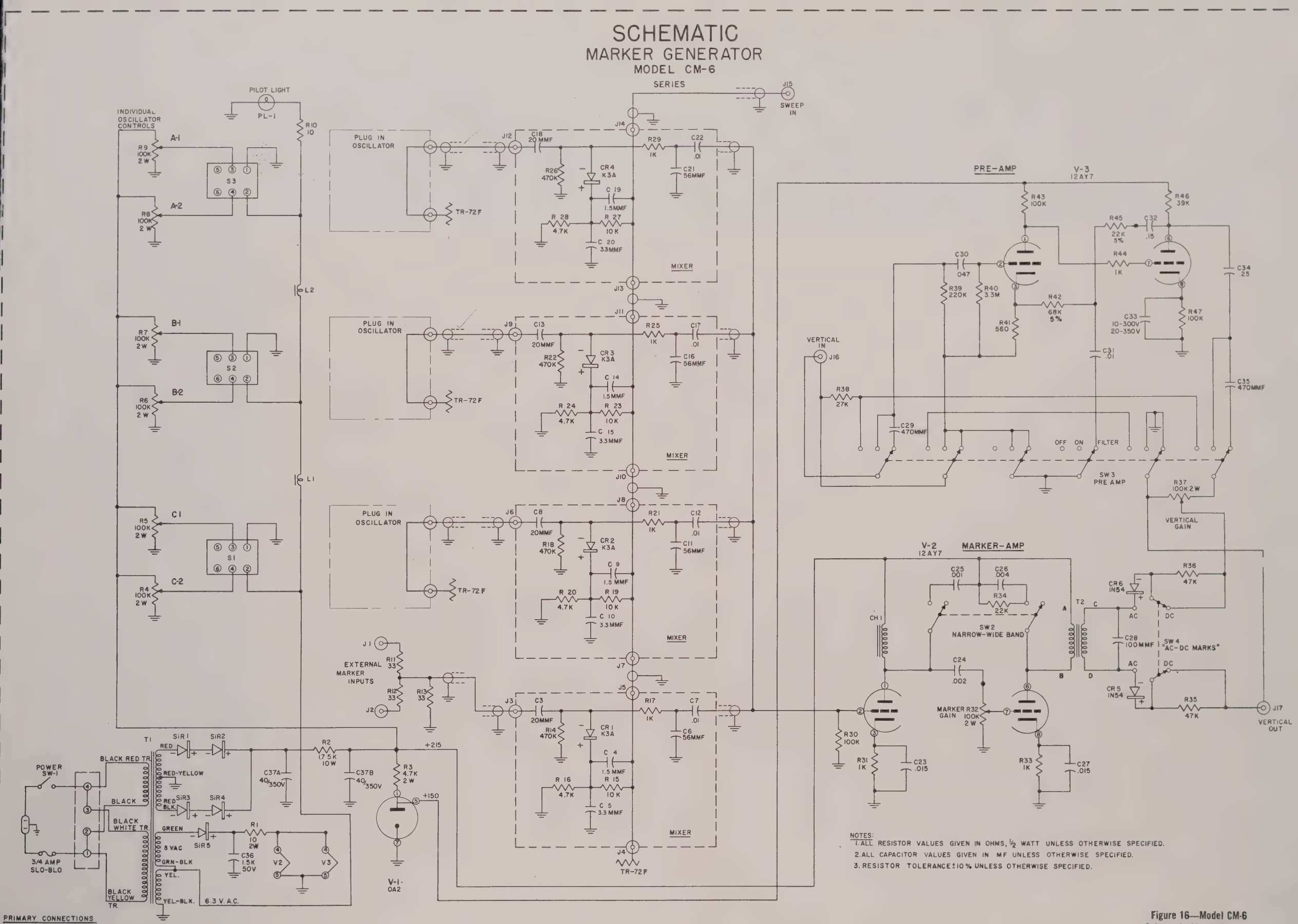



Figure 15

# SCHEMATIC
## MARKER GENERATOR
### MODEL CM-6

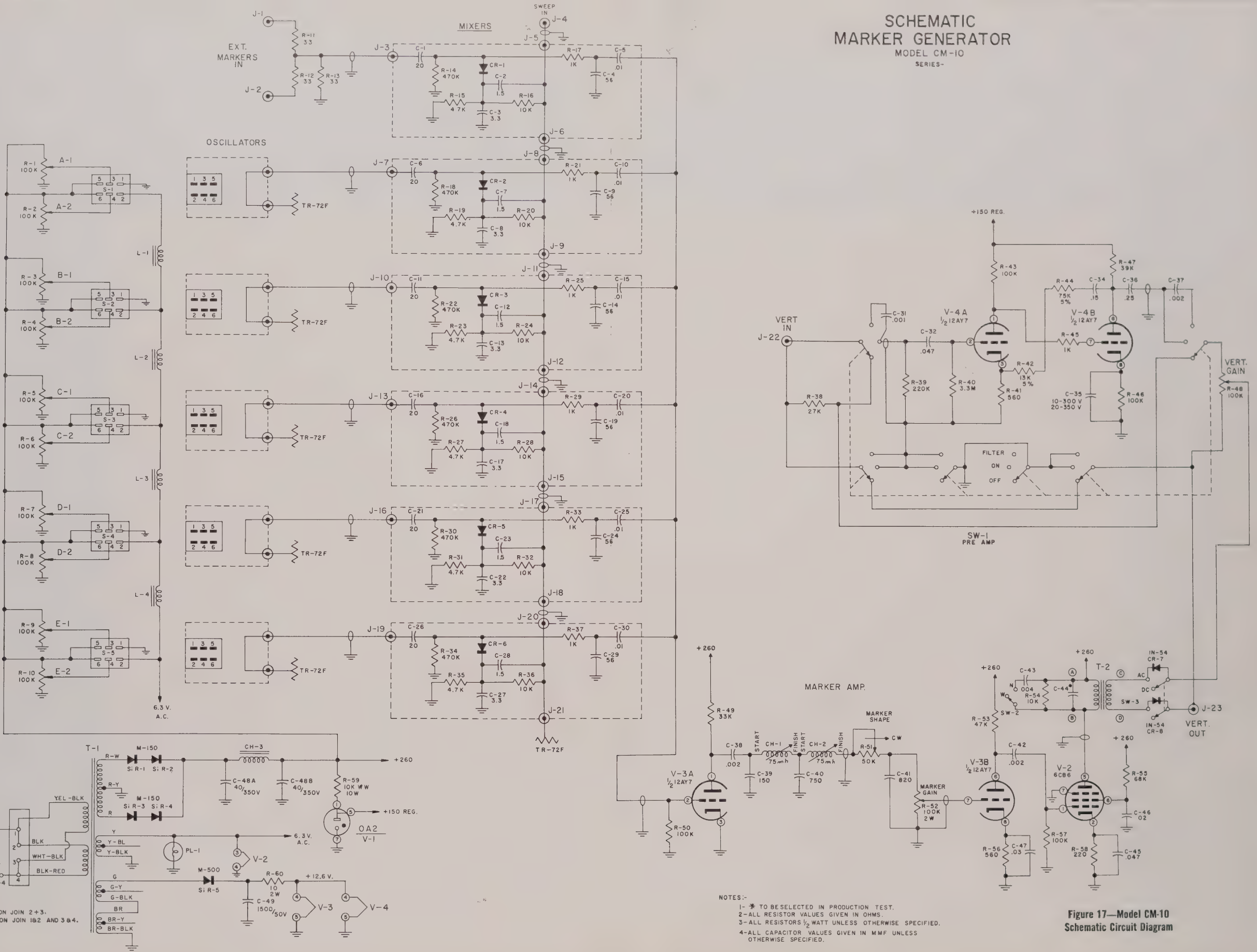



NOTES:
1. ALL RESISTOR VALUES GIVEN IN OHMS, ½ WATT UNLESS OTHERWISE SPECIFIED.
2. ALL CAPACITOR VALUES GIVEN IN MF UNLESS OTHERWISE SPECIFIED.
3. RESISTOR TOLERANCE ±10% UNLESS OTHERWISE SPECIFIED.

DWG. NO. 881-307-A

**Figure 16—Model CM-6 Schematic Circuit Diagram**

# SCHEMATIC
# MARKER GENERATOR
MODEL CM-10
SERIES-




NOTES:-
1- * TO BE SELECTED IN PRODUCTION TEST.
2- ALL RESISTOR VALUES GIVEN IN OHMS.
3- ALL RESISTORS 1/2 WATT UNLESS OTHERWISE SPECIFIED.
4- ALL CAPACITOR VALUES GIVEN IN MMF UNLESS OTHERWISE SPECIFIED.

**Figure 17—Model CM-10 Schematic Circuit Diagram**

DRAWING NO 861-336-1

JERROLD

# Technical Newsletter

Published by JERROLD ELECTRONICS CORPORATION, Philadelphia, Pa.
INDUSTRIAL PRODUCTS DIVISION

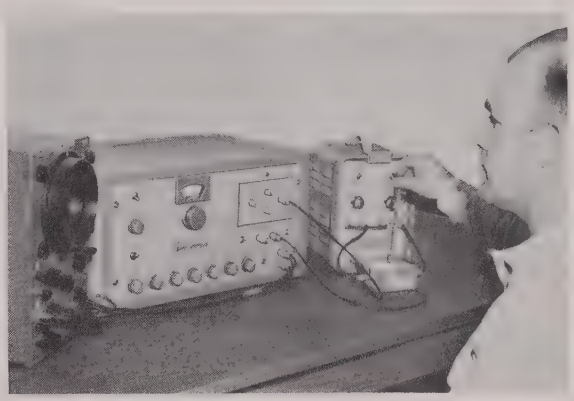
Volume 1, Number 1 Copyright © 1960. Jerrold Electronics Corporation, Philadelphia 32, Pa. March, 1958

# MEASUREMENT TECHNIQUES USING A COAXIAL SWITCH

**Since the genesis of the electronic age, hundreds of measurements have relied upon a system of comparison. In this system the desired information is obtained from the unit to be measured, and then compared to a known reference. This technique is capable of considerable accuracy in that it provides for the virtual elimination of test "set-up" errors.**

FIGURE 1 – ELECTRONIC COAXIAL SWITCH

An important disadvantage to the comparison method of test, however, is that it is generally quite time consuming. Manually operated coaxial switches have provided some degree of improvement insofar as the time consuming criticism is concerned, and electronic "hard-tube" switching has considerably improved measurements in the video and audio spectrums. Since electronic "Hard-tube" switching devices are limited to the video spectrum of 10 mc. and below, they are not applicable to direct R.F. measurements.



Indirect R.F. measurements can be accomplished however, by using the video switch to display alternately, the output of two R.F. detectors; one installed in the test circuit and the other in the reference circuit. Although this method is useful, its accuracy is limited by the requirement that the characteristics of the detectors and amplifiers must be balanced and precisely matched.

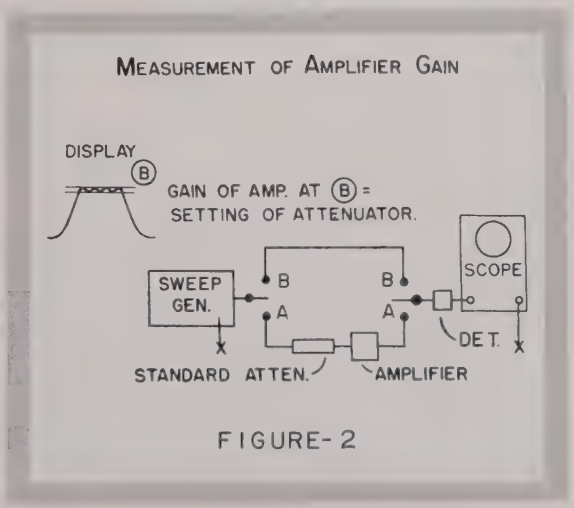



FIGURE-2

The Jerrold Coaxial Switcher, Model FD-30, was specifically designed to apply the comparison method of measurement in the frequency range from D.C. to 250 mc., without the need for costly, time consuming test "set-ups" and "take-downs", and without the problems of detector and amplifier balancing.

Figure 2 illustrates the use of an FD-30 in measuring the gain of an R.F. amplifier by the comparison technique. In switch position A the generator signal is fed to the amplifier to be tested through a variable attenuator and the amplifier output is connected to an R.F. detector. In switch position B, the generator signal is fed to the same R.F. detector through both switches. The Model FD-30 switches between positions A and B at a rate adjustable to 10, 15 or 30 cycles per second, so that the two detected outputs are seen super-imposed on the scope. Measurement is accomplished by adjusting the attenuator until the two traces coincide showing equality of the detector inputs in positions A and B. At this point, the gain of the amplifier equals the insertion loss of the attenuator.

The accuracy of the measurement is unaffected by variations in the generator's output, variations in the oscilloscope gain settings, or by the frequency and square law characteristics of the detector. Neglecting the attenuator accuracy, the only errors in the system are; VSWR or loss differences in paths A and B, and the "cross-talk" between the A and B switch sections.

To keep these errors at a minimum, the Model FD-30 was designed to provide for a VSWR of less than 1.1 up to 250 mc. The insertion loss of the switch elements is well below 0.5 db, and the "cross-talk" path has been designed not to exceed the coupling provided by 1.0 uuf or capacity, into a 50 ohm load at 250 mc. These specifications provide for accuracy capabilities - without refinement efforts - in the order of $\pm$ 0.2 db.

Improved accuracy can be accomplished, in any test procedure using the Model FD-30, by balancing the A and B paths for correction of the VSWR and insertion loss differences. This is accomplished by observing the oscilloscope presentation with the item to be tested replaced by a jumper, but utilizing all other cables and fittings. Attenuation differences in paths A and B can be corrected by the insertion of small value, fixed pads.

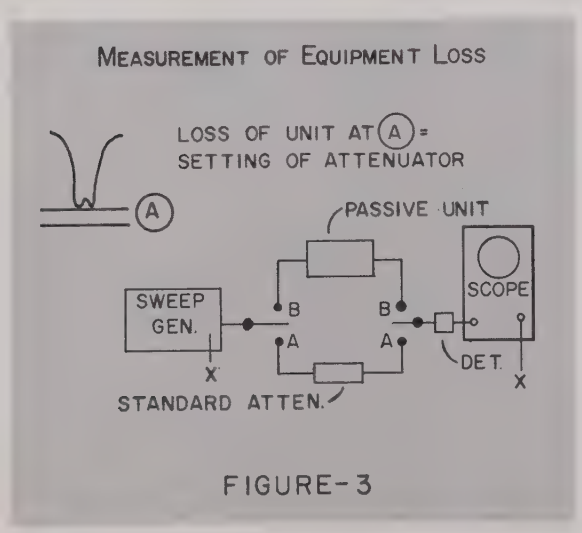



FIGURE-3

Illustrated in Figure 3 is a simple method of measuring the attenuation characteristics of filters, traps, coaxial cable and other passive networks. Here, the loss of the circuit is determined by inserting attenuation in the variable attenuator until the detected outputs of paths A and B are equal, as observed on the oscilloscope.

For circuits with losses as great as 40 db it may be necessary to use an oscilloscope preamplifier to improve the input sensitivity of the oscilloscope. The Jerrold Oscilloscope Preamplifier, Model SPR-100, is excellent for this purpose. The unit features a built-in detector that is flat from 500 kc. to 250 mc., an internal marker insertion amplifier-mixer. and an audio amplifier with a minimum gain of 40 db and good 60 cycle square wave response.

Fig. 4 illustrates a well-known and very useful sweep-frequency method of determining VSWR. A section of standard cable or "delay line" is connected to the output of a sweep-frequency generator, with a detector bridged across the junction. The detector supplies the scope with an input corresponding to the RF voltage at the line input.

Under these conditions the detector input voltage is the resultant of two components. The first component is E(m), the main wave, which is the only voltage when the delay line is terminated at the other end in its characteristic impedance. With a well-designed sweep this voltage is constant with frequency, and the scope displays a horizontal line.

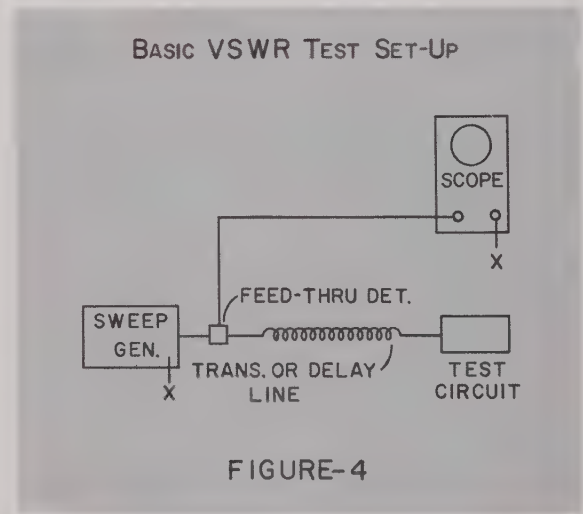



FIGURE-4

The other component is E(r), the reflected wave; this is present only when the far end of the line is mismatched, and its amplitude measures the degree of mismatch. When E(r) is in phase with E(m) they add to produce a voltage maximum; when it is 180° out of phase they subtract to a minimum. Due to the time delay in the line the phase of E(r) changes as the sweep frequency changes, so it alternately adds to and subtracts from E(m), producing a ripple pattern on the scope. At frequencies where the match is good, the reflection is small, and the ripple pattern has small amplitude; where the match is poor, there is a large reflection and the ripple has large amplitude.

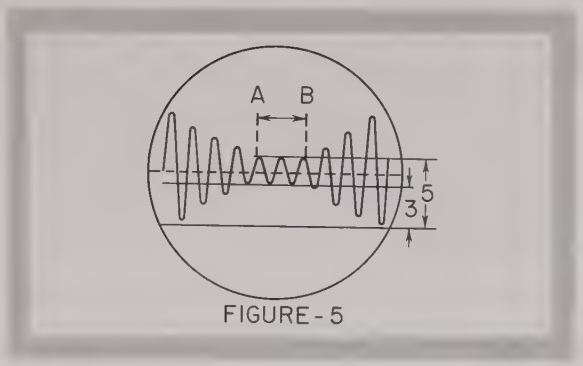


FIGURE-5

Figure 5 illustrates a typical scope display, showing the ripple pattern caused by a band-pass circuit which has a fair match between frequencies A and B and a poor match elsewhere. The horizontal dashed line shows where the trace would fall for a terminated line (E(m) only).

There are several well known methods of obtaining a VSWR measurement using this type of display. A common, but somewhat inaccurate method is to measure the relative height of Emax, and Emin as indicated on the oscilloscope. In Figure 5 Emax is indicated as 5 divisions and Emin as 3 divisions over the frequency range A-B.

***Since*** $\text{VSWR} = \frac{\text{Emax}}{\text{Emin}}$ ***Then*** $\text{VSWR} = 5/3 = 1.66$

The use of the divisions on the oscilloscope screen for a true voltage relationship does not allow for the square law characteristics of the detector. This would not normally be a large source of error when the VSWR approaches unity, but become an increasing source of error as the VSWR increases. Also, as the VSWR approaches unity it becomes increasingly difficult to use the oscilloscope face as a measurement of Emax and Emin, as these voltages are nearly identical and the difference between them is relatively small. The use of additional oscilloscope gain is limited by the need for the zero base line, and therefore, this method will not provide accurate VSWR measurements of small ratios.

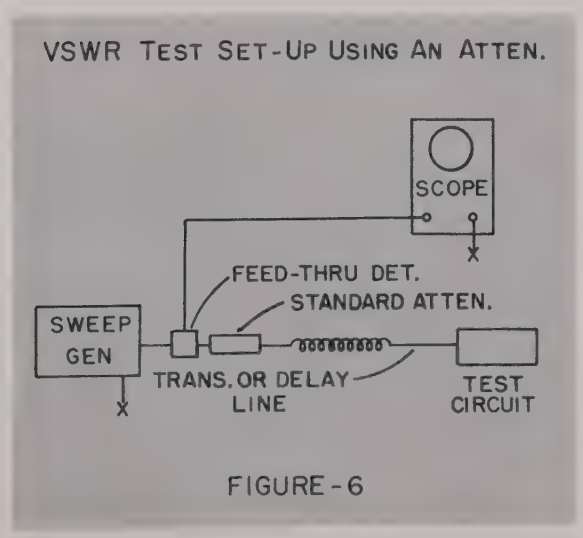


FIGURE-6

A more accurate method of VSWR measurements utilizes a calibrated attenuator as shown in Figure 6. First, all attenuation is removed from the attenuator and the unit to be measured is connected to the end of the transmission line. The height of the reflected voltage as shown on the oscilloscope is recorded. The unit is now removed from the line and the line is left opened to provide 100% reflection. Attenuation is inserted in the attenuator until the height of the reflected voltage equals the reflected voltage from the circuit under test, as previously recorded. The amount of attenuation inserted, now provides us with the relationship between E(m) and E(r). If 10 db insertion was required in the attenuator we may then say that the "return loss" of the circuit is 20 db, since the reflected voltage has to make two trips through the attenuator. If a VSWR relationship is required, we have:

$$\text{VSWR} = \frac{E(m) + E(r)}{E(m) - E(r)}$$

AND: E(r) = 20db below E(m)
OR: E(r) = 1/10 E(m)
AND: E(m) + E(r) = 1+ .1 or 1.1
E(m) - E(r) = 1 - .1 or .9

THEREFORE: $\text{VSWR} = \frac{1.1}{.9}$ **or 1.23**

NOTE: See nomograph of VSWR, Figure 8

The advantage of this method over the previous method is that the result is not affected by the square law characteristics of the detector, and all of the oscilloscope gain may be utilized since we no longer require the base line to determine the relationship between E(m) and E(r).

The delay line losses are also of no consequence (except for sensitivity) in that the calbriated attenuator was adjusted to supply the same reduction of E(r) as the test unit provided and the line losses were identical in both indications.

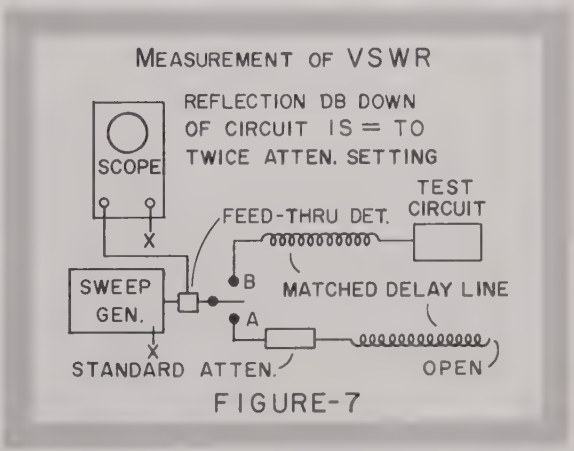


FIGURE-7

To further simplify the test procedure, the Model FD-30 may be used to simultaneously display both the reference and the test circuit. This basic test set-up is shown in Figure 7. All that is required in this procedure is that we provide identical transmission lines for the reference and test circuit. This can be accomplished by cutting both lines from the same reel of cable while making the test circuit line slightly longer than the reference circuit line. The lines can be adjusted for identical characteristics by leaving both unterminated while observing them simultaneously on an oscilloscope, and cutting the line that displays the greater electrical length to match the shorter line.

Figures 9 and 10 show two other interesting applications for the Jerrold Model FD-30.

If measurement is required of match that exhibits return losses in excess of 35 db, improvement in display indication can be obtained by utilizing a high pass audio filter between the detector and the oscilloscope. The filter is designed to pass the audio frequencies of the ripple voltage while providing 20-30 db of attenuation for the 60 cycle and lower frequency components of the E(m) signal. This feature emphasizes the E(r) component and thus faciltates the measurement. The required filter section is an integral part of the Model SPR-100 Oscilloscope Preamplifier. The filter can be switched in or out of the circuit as needed. Figure 11 illustrates the complete equipment package required for VSWR measurements.

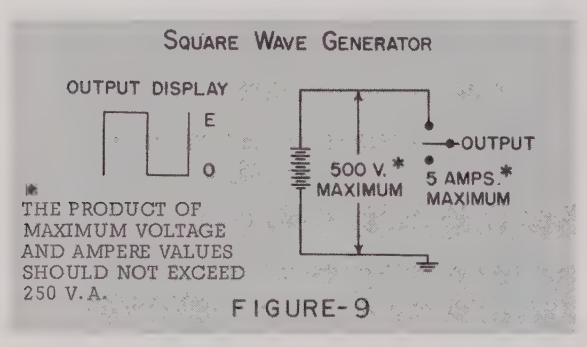



FIGURE-9

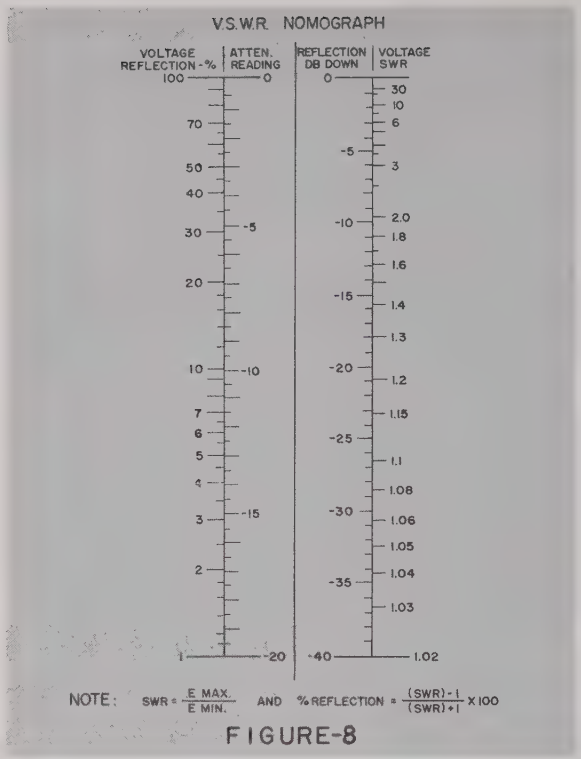

V.S.W.R. NOMOGRAPH
VOLTAGE REFLECTION - %
ATTEN. READING
REFLECTION DB DOWN
VOLTAGE SWR
NOTE: SWR = E MAX. / E MIN. AND % REFLECTION = (SWR)-1 / (SWR)+1 X 100

FIGURE-8

## CALIBRATING SCOPE DISPLAY OF DIODE OR TRANSISTOR CHARACTERISTICS

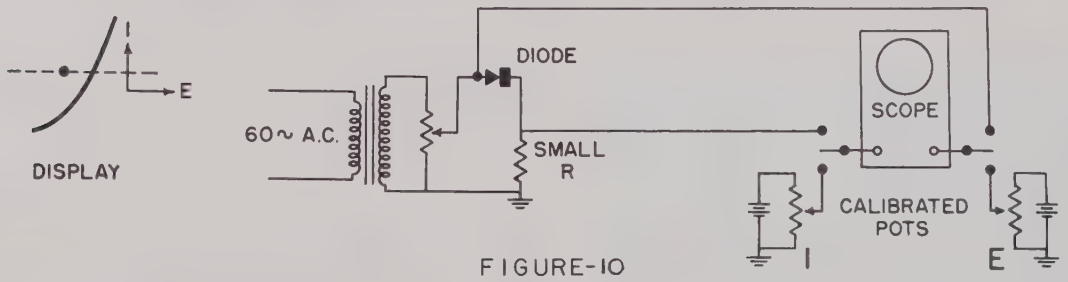
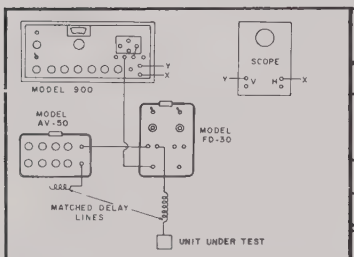



FIGURE-10




| FREQUENCY RANGE 200 KC to 250 mcs | | |
|---|---|---|
| MODEL | DESCRIPTION | PRICE |
| 900 | SWEEP GENERATOR (0.2 - 1,000 MCS) OSCILLOSCOPE PREAMPLIFIER (included in Sweep Generator) | $1260. |
| FD-30 | COAXIAL SWITCH | $ 250. |
| AV-50 | VARIABLE ATTENUATOR | $ 150. |
| | TOTAL COST | $1660. |

| FREQUENCY RANGE 12 mcs to 225mcs. | | |
|---|---|---|
| MODEL | DESCRIPTION | PRICE |
| 601 | SWEEP GENERATOR | $390. |
| FD-30 | COAXIAL SWITCH | $250. |
| SPR-100 | OCILLOSCOPE PREAMPLIFIER | $200. |
| AV-50 | VARIABLE ATTENUATOR | $150. |
| | TOTAL COST | $990. |

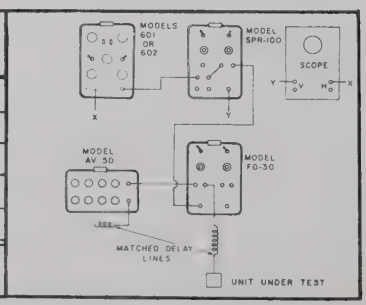



FIGURE-11

*Prices Subject To Change Without Notice*

**Export:** Rocke International, New York 16, New York. Cable—ARLAB. **Canada:** Jerrold Electronics (Canada) Ltd., Toronto, Canada.

JERROLD

# Technical Newsletter

Published by JERROLD ELECTRONICS CORPORATION, Philadelphia, Pa.
INDUSTRIAL PRODUCTS DIVISION

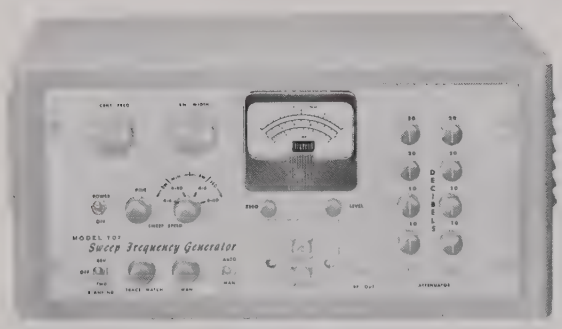
Volume 1, Number 2 COPYRIGHT © 1959 BY JERROLD ELECTRONICS CORP., PHILADELPHIA 32, PA. August, 1958

# Wide Band Impedance Measurement

## With a Sweep-Frequency Generator and Delay Line

by KEN SIMONS
*Chief Test Equipment Engineer*
*Jerrold Electronics Corporation*

**The Sweep-Frequency, Delay Line method of testing was first used in the alignment and test of TV Broadcasting Antennas. Since then it has been widely used in the TV industry where impedance match over a wide band of frequencies was important. This article discusses the basic principles involved, and describes some techniques that allow an increase in measurement accuracy.**




MODEL 707
Sweep Frequency Generator

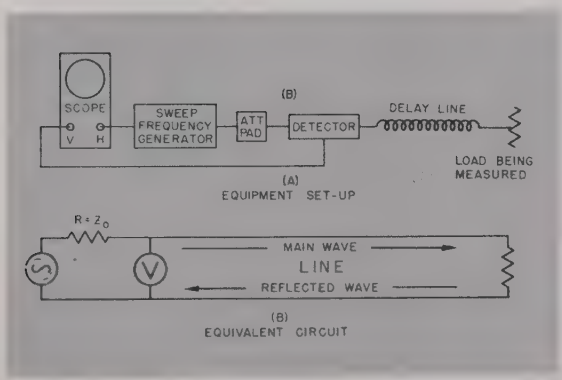



FIGURE 1

### Equipment Set-Up

The basic equipment test set-up for wide-band impedance measurements is shown in Fig. 1 (a). An attenuator pad is generally connected between the sweep and the detector to isolate the generator from the varying impedance, and to ensure that the detector is fed from a well matched source. The detector is needed when the sweep covers a frequency range above the upper response limit of the scope. The high-frequency voltage is rectified, and the detected output, a slowly-varying voltage representing the envelope of the high-frequency input, is applied to the oscilloscope.

The equivalent circuit of the test set-up is shown in Fig. 1 (b). A voltmeter (the detector and scope) measures the voltage at the junction between a matched source and a transmission line with a load on its far end. The voltage at this junction may be considered as being due to the sum of two waves: the *main wave* energy coming out of the sweep and going down the line, and the *reflected wave;* energy which has travelled down the line, been reflected from the load, and come back up the line.

# The Delay Line

## Terminated Line

With a well-designed sweep, the main-wave is made approximately constant with frequency. When the delay line is terminated in a resistance equal to its characteristic impedance, there is no reflected wave, so the scope shows a constant voltage. Fig. 2 illustrates a typical pattern. On the forward trace the output stays constant as the frequency changes across the sweep band. The reverse trace is "blanked"; the sweep output being keyed to zero to provide a reference line showing where zero output is.

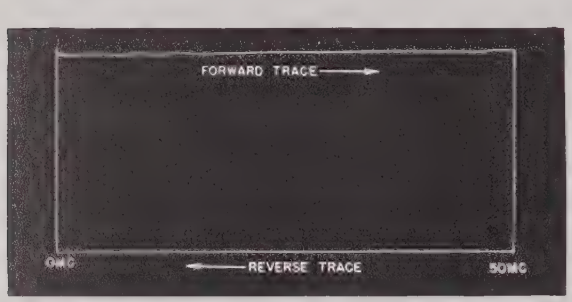



FIGURE 2

## Open Circuit Line

When nothing is connected to the end of the line this "open circuit" cannot dissipate, so all the energy striking it is reflected. When the line loss is small, the reflected wave at the detector has essentially the same amplitude as the main wave, and the total voltage there depends only on their phase relation. When they are in phase, they add to produce a voltage twice the "main wave only" condition. When they are 180° out of phase, they cancel to produce zero output. Fig. 3 illustrates the pattern with a generator sweeping 0 to 50 mc, and an open-circuited delay line ½ wavelength long at 50 mc. The reflected wave *at the open circuit* is in phase, at all frequencies, with the main wave. Near zero frequency the effective length of the line is zero so the two waves are in phase and add at the detector, producing a maximum. At 25 mc, the line is ¼ wave-length long; the main wave shifts 90° in phase as it travels down the line to the end; and the reflected wave shifts another 90° on its way back, so the two components are 180° out of phase at the detector, giving a minimum at that frequency. At 50 mc, the reflected wave has travelled a full wavelength (½ wave down, ½ wave back) by the time it gets back to the detector so the components add to a maximum.

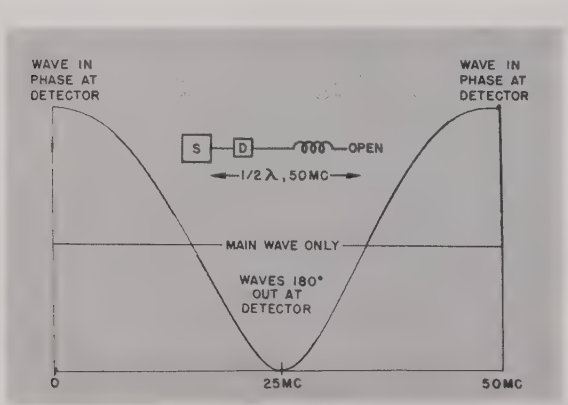



FIGURE 3

## Short Circuit Line

Fig. 4 illustrates the pattern obtained by connecting a short circuit across the end of the line. This reverses the phase of the reflected wave so that the two components are 180° out at the short, and produce zero voltage there. At the detector, this produces a minimum at zero frequency (effective line length 0), a maximum at 25 mc., (reflected wave shifted 180° having travelled ½ wavelength extra), and a minimum again at 50 mc.

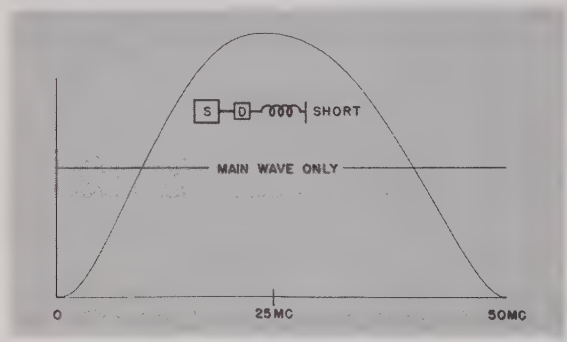



FIGURE 4

## Longer Delay Lines

Fig. 5 shows the effect of doubling the length of the line, keeping the sweep-width the same. With an open-circuit (a) there are maxima at 0 frequency (0 line length), 25 mc. (½ wavelength) and 50 mc (1 wavelength) and minima when the line length is ¼ and ¾ wavelength. A short-circuit termination gives the reverse pattern (b).

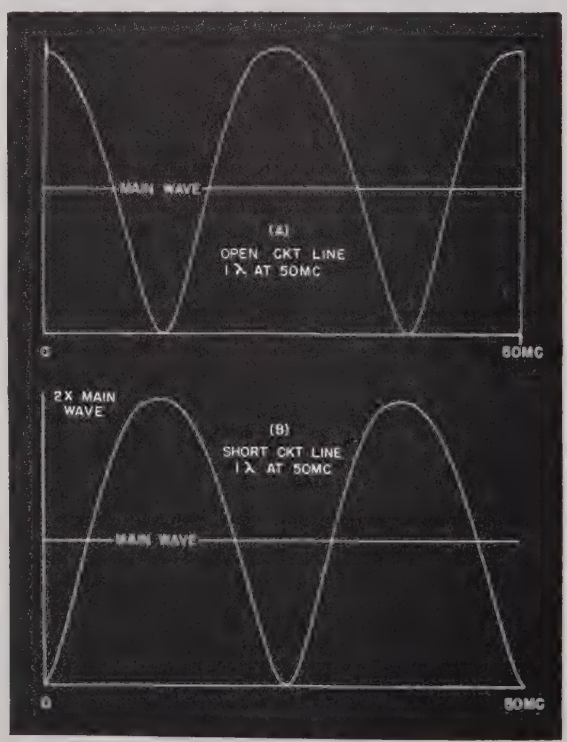



FIGURE 5

It is apparent that increasing the length of the delay line increases the number of ripples occuring in a given sweep-width. Specifically, the frequency change from one maximum (or minimum) to the next is equal to the frequency at which the line is one-half wavelength long. For a spacing of f mc between peaks, the line length in feet is $\frac{492\ d}{f}$ where d is the delay factor, the ratio of the speed with which a wave travels in the particular type cable, to the speed in air. d equals 0.67 for solid polyethylene insulated cables, and approximately 0.8 for polyfoam or other dielectrics with a high proportion of air insulation.

Since more ripples depict more clearly what happens in a given frequency band, in this respect, longer lines are preferred over shorter ones. One factor that limits the length that may be effectively used is the line attenuation. As the length is increased more and more energy is lost in the line, so the reflected signal, as it shows up back at the detector, gets increasingly weaker. Fig. 6 (a) shows the ripple pattern resulting with a short-circuit termination, a sweep from 50 to 100 mc., and a high-grade delay line for 5 mc. between peaks.

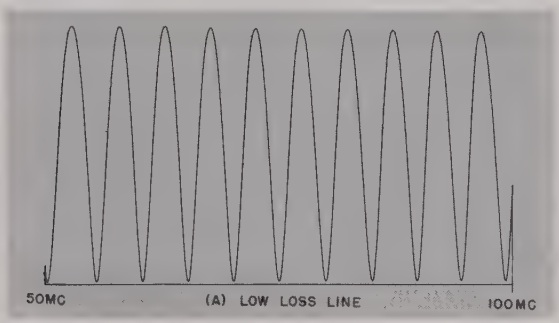



FIGURE 6a

The delay factor for this cable was 0.8, so its length, by the formula, would be $\frac{492 \times 0.8}{5}$ or 79 feet. Due to the greater length of this line, and the higher frequencies involved there is appreciable loss in the line. Although the reflected wave at the shorted end is equal to the main wave (reflection—100%), the main wave is stronger at the detector than it is at the far end, due to attenuation as it travels down the line, and the reflected wave is weaker at the detector than it is at the load, due to attenuation as it travels up the line from the load. This shows up on the ripple pattern in the fact that the minima do not quite go to zero, since the weakened reflected wave does not quite cancel the main wave.

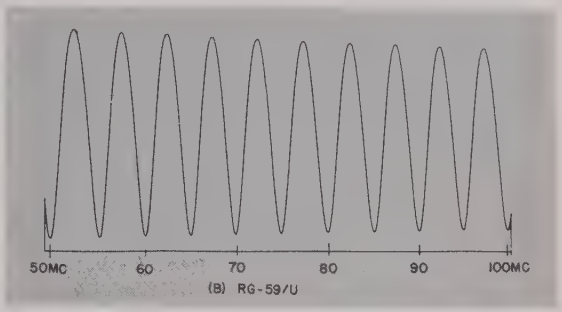
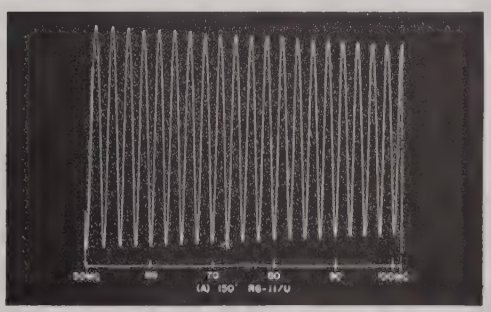



FIGURE 6b

Fig. 6 (b) shows the effect of using a cheaper cable (RG59/U) with higher loss. This line was cut for the same electrical length, but because its delay factor is 0.67, its physical length was shorter, 66 feet. Its greater attenuation shows up in the fact that the minima are still farther from zero. Notice that this effect is more pronounced at the high frequency end of the sweep, where the line loss is higher.

Fig. 7 (a) shows the ripple pattern from approximately 150' of RG11/U cable, shorted at the far end. Notice that the amplitude of the ripple pattern is about the same as that from the 66' piece of 59/U in Fig. 6 (b). The attenuation of RG11/U is about one-half that of RG59/U for a given length; so doubling the length just about cancels the attenuation improvement due to the larger cable. Fig. 7 (b) shows the pattern from about 500' of RG11/U showing the great reduction in ripple height at this length.



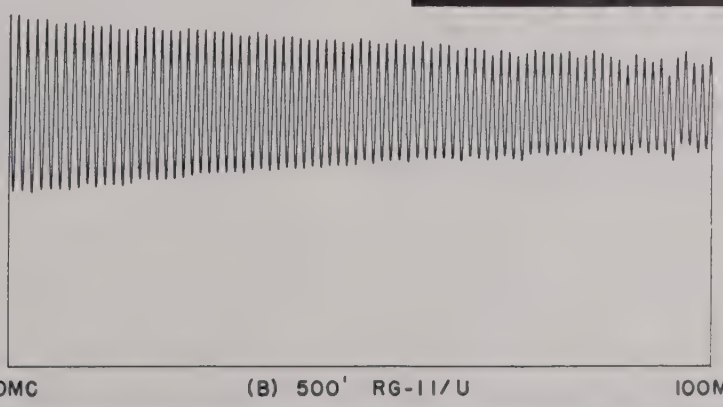




FIGURE 7

# Terminations

## Various Resistive Terminations

The wave reflected from a purely resistive termination is in phase with the main wave at the load for resistance values higher than Zo, or 180° out of phase for values lower than Zo, and the amplitude of the reflection increases as the resistance differs from Zo.

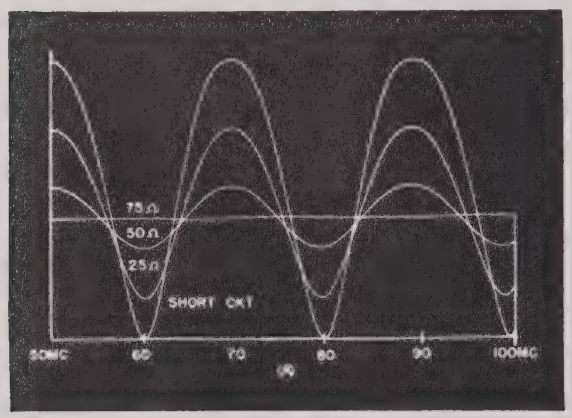



FIGURE 8a

Fig. 8 (a) shows superimposed the ripple patterns obtained with a low-loss 20 mc delay line, a sweep from 50 to 100 mc, and the indicated terminations. Note that all the patterns have minima at the same frequency as that with a shorted end.

> The impedance of a load giving a minimum at the same frequency as a short circuit is resistive and lower in resistance than Zo.

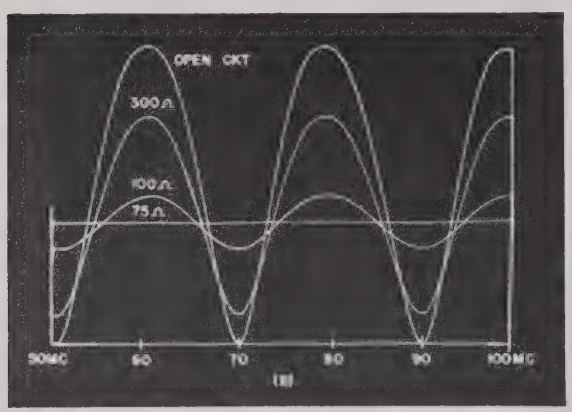



FIGURE 8b

Figure 8 (b) shows the patterns from several resistive terminations higher than Zo. They all have maxima at the "short-minimum" frequencies.

> The impedance of a load giving a maximum at the same frequency as a short-circuit minimum is resistive and higher in resistance than Zo.

## Purely Reactive Terminations

A pure reactance does not dissipate energy, so the wave is reflected from a purely reactive termination at full amplitude. Its phase is shifted depending on the magnitude of the reactance relative to Zo. Thus the ripple pattern obtained with purely reactive terminations has the same amplitude as with an open or short circuit, but the minima and maxima are shifted in frequency. Fig. 9 (a) shows the patterns obtained with two sizes of capacitors compared with open and short circuit patterns.

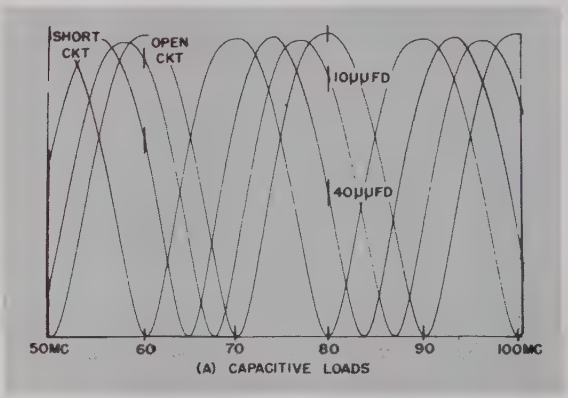



FIGURE 9a

The ripple pattern resulting from a capacitive termination has minima falling higher in frequency than those due to a short circuit, and lower than those due to an open circuit. This may be stated another way:

> If we mark the frequencies of short circuit minima (see marks on Fig. 9a) the marks will fall on down slopes with a capacitive load.

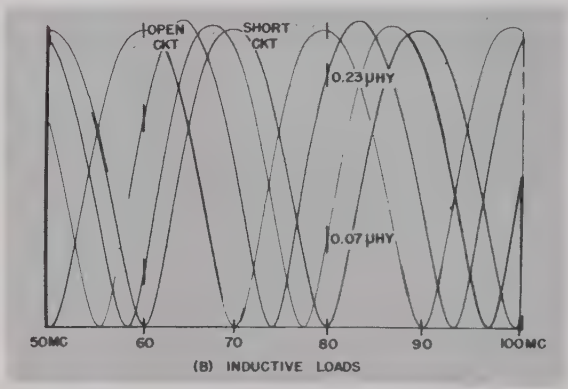



FIGURE 9b

Fig. 9 (b) compares the ripple patterns for two sizes of purely inductive terminations with those from an open and a short circuit.

> The ripple pattern from an inductive termination has minima falling above those from an open circuit, and below those from a short circuit. If we mark the frequencies of short circuit minima, the marks fall on upslopes for inductive terminations.

## Complex Terminations; Impedance Changes with Frequency

When a termination has both dissipation (resistance) and reactance the reflected wave is reduced in amplitude, and shifted in phase relative to the main wave. Correspondingly, the ripple pattern has a lower amplitude, and minima shifted in position compared with a short or open circuit. Fig. 10 (a) illustrates the ripple pattern obtained with a low-loss 5 mc delay line and a series RC termination. At higher frequencies, where the reactance is lower, the termination approaches a matched condition, and the ripple has lower amplitude. The nature of the termination can be seen more clearly if the ripple pattern is compared with that obtained with a short circuit, and that obtained with a matching resistance. Fig 10 (b) shows the three patterns superimposed. By observing the fact that down-slopes occur at shorted-minimum frequencies, and the ripple gets smaller at high frequencies, we could conclude that the load had the characteristics of a series RC circuit.

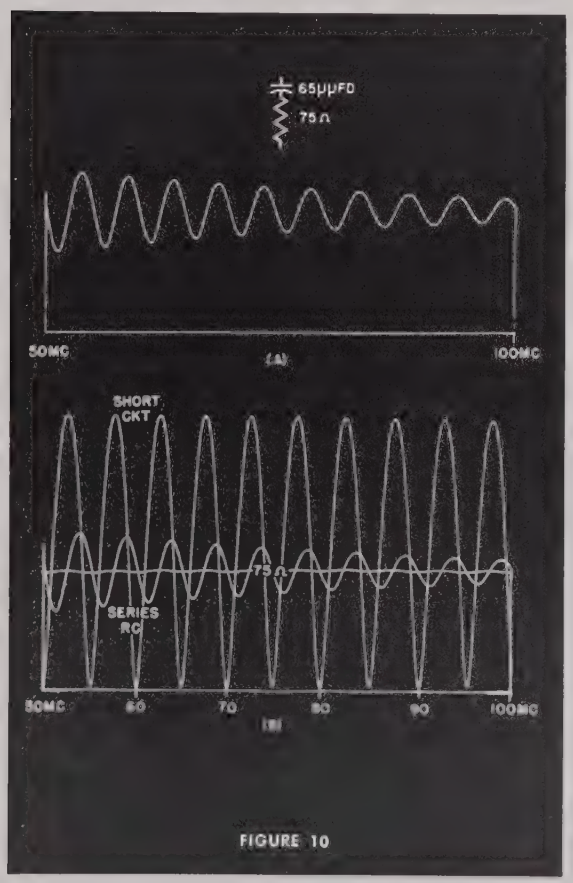



FIGURE 10

Fig. 11 (a) shows the ripple pattern with a 20 mc line and a series resonant circuit. Fig. 11 (b) shows the same pattern superimposed on a short circuit pattern and one from a matched resistor. By observing that its impedance is capacitive below the resonant frequency (down-slope at short-circuit minimum frequencies), matched at resonance (low ripple amplitude near 75 mc), and inductive above resonance (up-slope of shorted-minimum) we could deduce that it was a series RLC circuit with R equal to Zo.

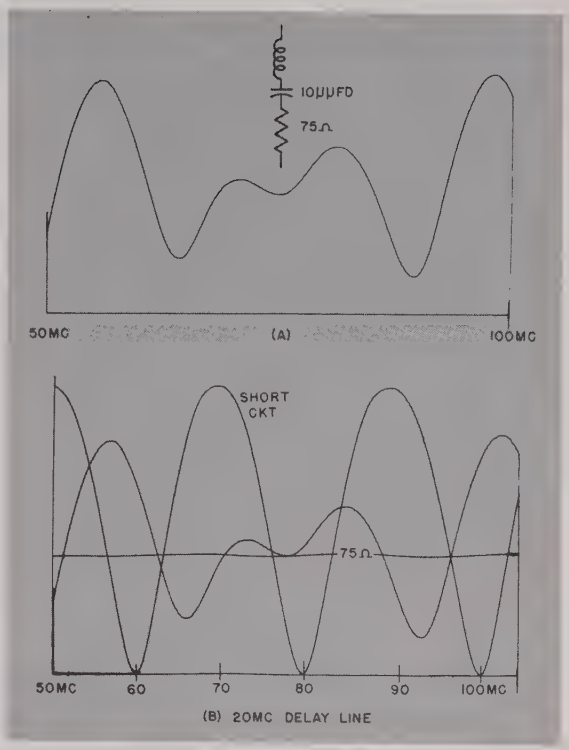



FIGURE 11

Fig 12 (a) and (b) show the ripple patterns obtained under the same conditions as Fig. 11 (b) but with 5 mc and 2.5 mc delay lines respectively. They show how increased length in the delay line depicts the impedance characteristic in more detail.

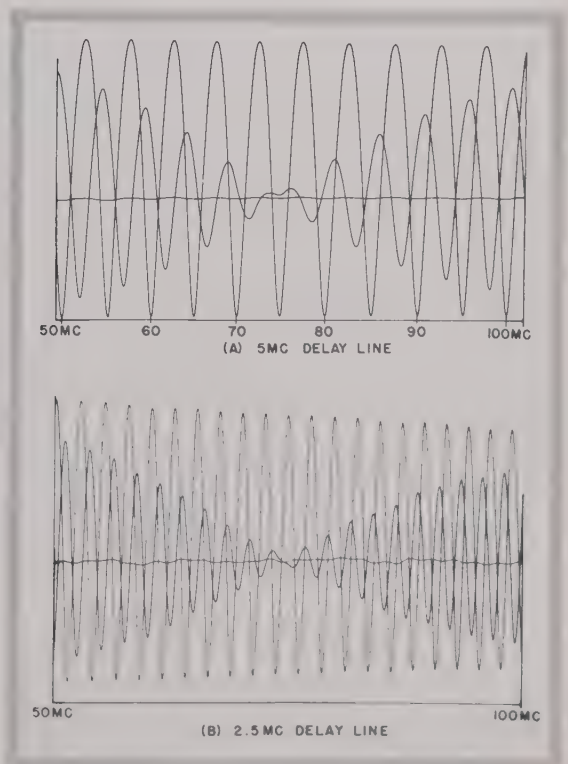



FIGURE 12

# Determining Reflection Coefficient and VSWR

There is a temptation, in using the delay-line technique, to assume that the VSWR of the load is found by taking the ratio of minimum to maximum of the ripple pattern displayed on the scope. Two factors, the loss of the delay line and the non-linearity of the detector, make this procedure quite inaccurate. A more accurate procedure that eliminates the effect of the line loss and minimizes detector non-linearity is to compare the amplitude of the ripple pattern from the unknown with that from a short- or open-circuit.

Let A be the peak-to-peak amplitude of the ripple pattern obtained at the frequency of interest with the line shorted. Let B be the peak-to-peak amplitude of the ripple pattern from the unknown at this frequency.

Then the reflection coefficient of the unknown (ratio of main wave to reflected wave): K equals $\frac{B}{A}$

*Percent reflection* equals 100K

The *Return Loss* (reflection coefficient as a DB ratio) is $20 \log_{10} \frac{1}{K}$

The VSWR (ratio of max voltage at load to min. voltage is $\frac{1 + K}{1 - K}$ These various ways of expressing the degree of mismatch are conveniently related by the nomogram shown in Fig. 13.

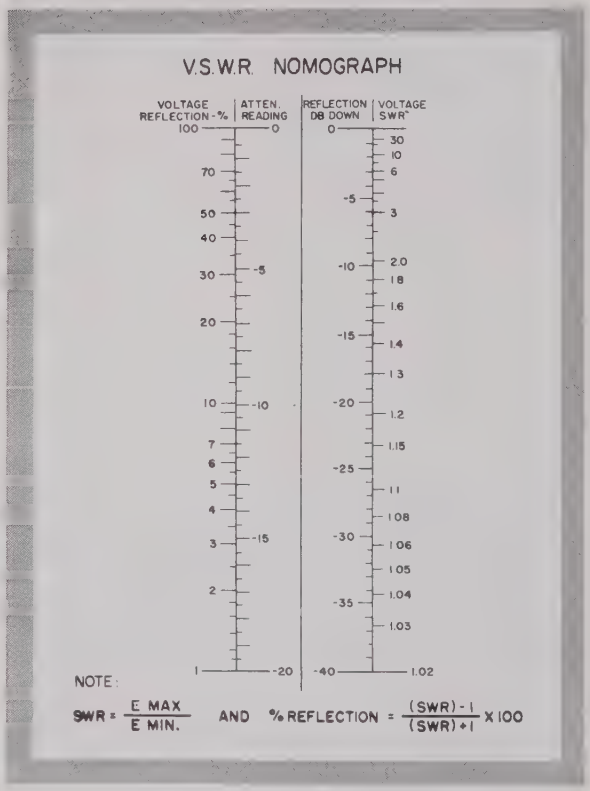

V.S.W.R. NOMOGRAPH
VOLTAGE REFLECTION - %
ATTEN. READING
REFLECTION DB DOWN
VOLTAGE SWR
NOTE:
SWR = E MAX / E MIN. AND % REFLECTION = (SWR) - 1 / (SWR) + 1 X 100

FIGURE 13

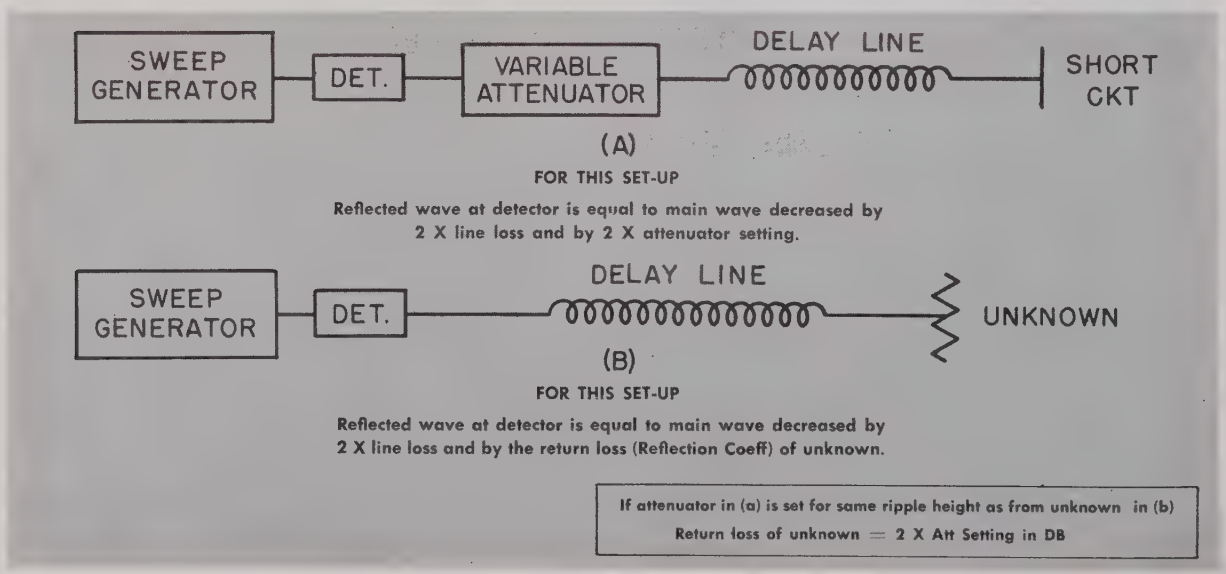

SWEEP GENERATOR
DET.
VARIABLE ATTENUATOR
DELAY LINE
SHORT CKT
(A)
FOR THIS SET-UP
Reflected wave at detector is equal to main wave decreased by 2 X line loss and by 2 X attenuator setting.
SWEEP GENERATOR
DET.
DELAY LINE
UNKNOWN
(B)
FOR THIS SET-UP
Reflected wave at detector is equal to main wave decreased by 2 X line loss and by the return loss (Reflection Coeff) of unknown.
If attenuator in (a) is set for same ripple height as from unknown in (b)
Return loss of unknown = 2 X Att Setting in DB

FIGURE 14

A more accurate, and generally more convenient way of determining the reflection coefficient of a load is to compare the height of its ripple pattern with the height of the pattern obtained with a short at the far end of the line, and a variable attenuator inserted between the line and the detector. Fig 14 illustrates the calibration set-up. By putting the attenuator at the detector end of the line the ripple height is determined primarily by its attenuation, and its impedance match is less important than if it were connected at the load end of the delay line.

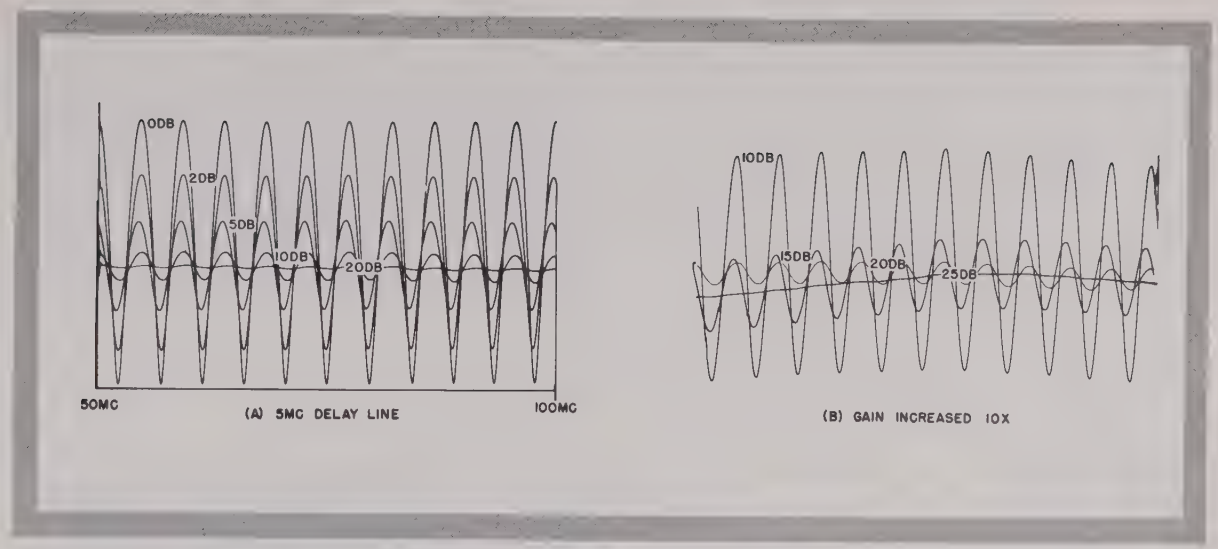



FIGURE 15

Fig. 15 illustrates the ripple patterns obtained at various settings of an attenuator connected in this way. Note that the ripple height with, for example, 5 db set on the attenuator, is what would be seen from a load having a return loss of twice this many db (e. g. 10 db). Fig. 15 (b) was made by increasing the vertical gain 10X compared with (a). Note that the ripple corresponding to a return loss of 40 db (VSWR 1.02) is readily seen. Also, note the irregularities in the way these traces match up. These are due to very small errors in the impedance match and delay of the attenuator used.

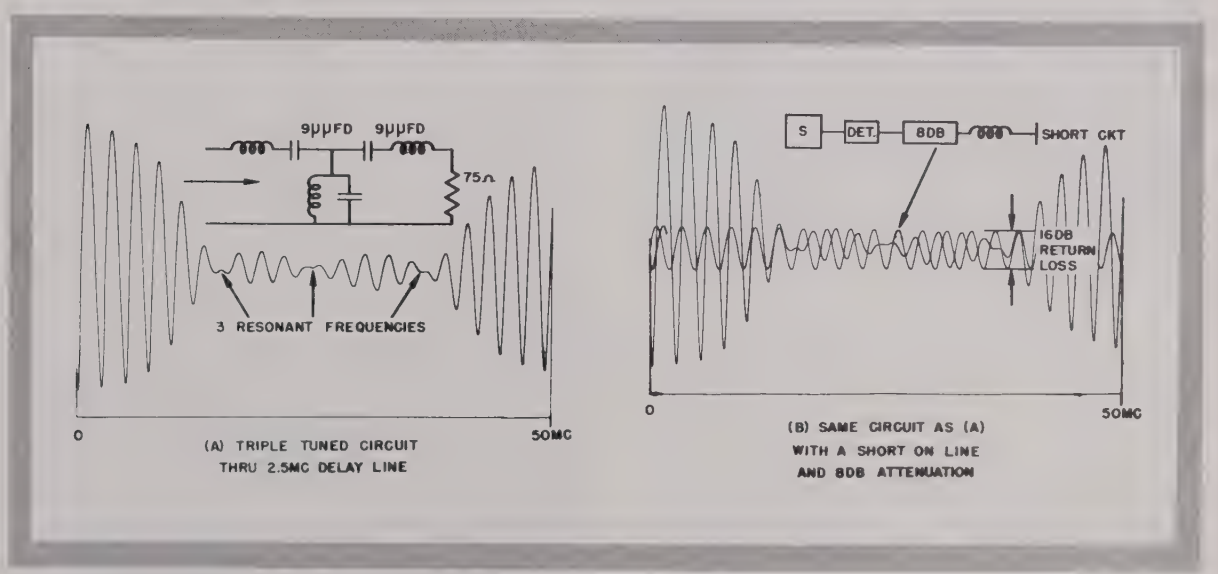



FIGURE 16

Fig. 16 illustrates how the technique is used in determining return loss. A triple-tuned band-pass filter was connected to the end of a 2.5 mc delay line, giving the ripple pattern of Fig. 16 (a). Note the three frequencies of best match. To determine the maximum reflection within the pass band, the ripple from a shorted line through a variable attenuator was set to the same height as the maximum ripple in the pass band. The attenuator read 8 db. indicating a maximum return loss for this filter of 16 db. The two patterns are shown superimposed in Fig. 16 (b).

# Irregularities in the Delay Line

With reasonable sweep output level and scope gain, it is not difficult to see ripples corresponding to 40 or even 50 db return loss. It is not generally possible to make measurements with this much accuracy. The limiting factor is not gain, but the uniformity of the delay time. Many commercial coaxial cables have a degree of nonuniformity that results in appreciable reflections from within the cable, even when terminated with the best possible load. A poor cable may have reflections with a return loss of as little as 20 db. Most cables run at or above 30 db., and only an exceptionally uniform cable has internal reflections more than 40 db down. To illustrate the extremes that may be encountered several traces were made. Fig. 17 (a) shows the ripple pattern of a 5 mc delay line of exceptional uniformity terminated in a load that closely matches its impedance. It was possible to see variations in the pattern only by increasing the vertical gain 10X (Fig. 17b). Comparing the ripple pattern with the calibration pattern corresponding to 40 db return loss, it can be seen that the combined return loss of cable and termination is decidedly better than 40 db over the whole frequency range of the sweep. For contrast, Fig. 17 (c) shows the trace with a very poor piece of cable. Even with the best termination possible, its variation is more than 10X greater than the other cable. (compare with 17a).

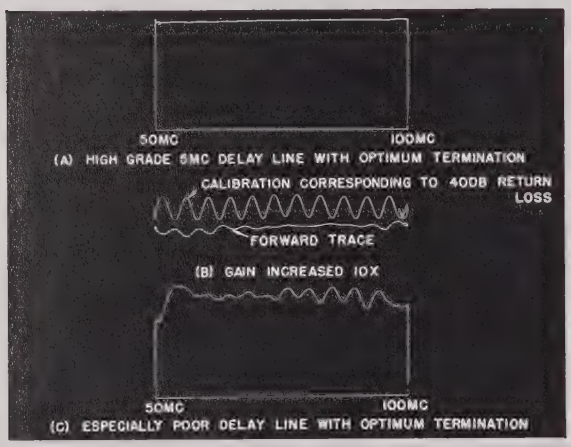



FIGURE 17

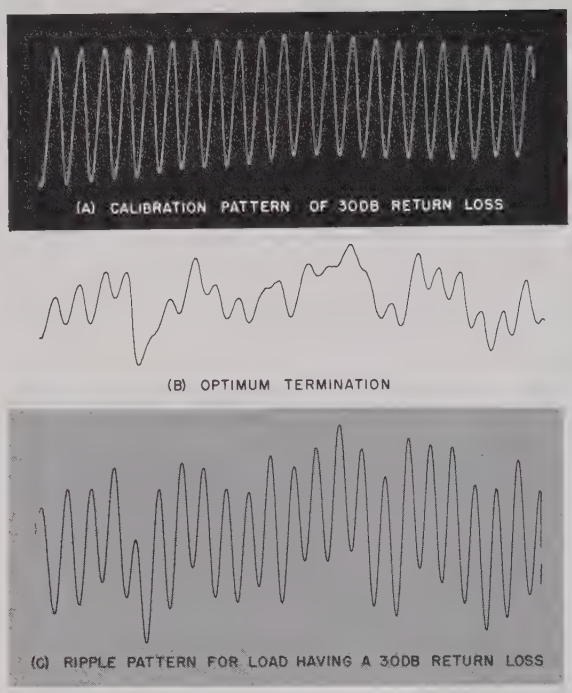



FIGURE 18

Fig. 18 is an illustration of the way in which irregularities in the delay line can make accurate measurement difficult. (a) shows the ripple pattern amplitude under a calibration condition corresponding to 30 db return loss (high vertical gain was used). (b) shows the pattern with termination adjusted for minimum ripple. This line, a 150' length of RG11/U, has internal reflections a little more than 30 db down. (c) shows the pattern resulting when a load having a return loss of 30 db was connected to the end of this line. It can be seen that the line irregularities prevent accurate display of the load characteristics. A further hazard is that the internal reflections probably indicate a considerable variation in characteristic impedance of one section of the cable as compared with another. Thus, the impedance which will match the far end depends on just where the cable is cut.

# A Note on Comparison Technique

The patterns illustrating this article were recorded with a Moseley X-Y recorder, using a Jerrold Model 707 Precision Sweep Generator, which has sweep speeds adaptable to use either with a recorder or a normal oscilloscope. Where patterns are shown superimposed, they were made by simply changing the load connections without touching the sweep settings and recording the second trace on top of the first. This is, of course, not possible with a scope, but the same convenience of simultaneous presentation for calibration can be obtained by using a Jerrold Model FD-30 Coaxial Switch. For Delay Line work, it is connected as shown in Fig. 19, superimposing a calibration pattern from a second delay line on the pattern of the load being adjusted or measured.

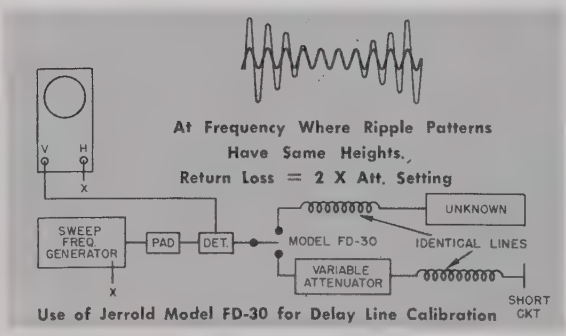



FIGURE 19

**NOTE — Additional copies of this paper and engineering assistance on the application of Measurements By Comparison may be obtained by contacting the Industrial Products Division, Jerrold Electronics Corporation, 15th and Lehigh, Philadelphia 32, Pa. Phone: BAldwin 6-3456 — Ext. 211.**

EDITORIAL STAFF: *Managing Editor* - C. C. Cooley, Jr. — *Technical Copy Editor* - A. Kirkaldy, Jr. — *Art Director* - R. S. Reisner, Jr.

**Export:** Rocke International, New York 16, New York. Cable—ARLAB. **Canada:** Jerrold Electronics (Canada) Ltd., Toronto, Canada.

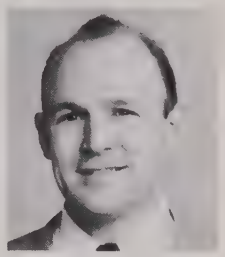
JERROLD

# Technical Newsletter

Published by JERROLD ELECTRONICS CORPORATION, Philadelphia, Pa.
INDUSTRIAL PRODUCTS DIVISION

Volume 1, Number 3 COPYRIGHT © 1959 BY JERROLD ELECTRONICS CORP., PHILADELPHIA 32, PA. October, 1959

## Additional Sweep-Frequency Impedance Measuring Techniques



By KEN SIMONS, Chief Engineer, Jerrold Electronics Corp.

**The use of a Delay Line with a Sweep-Frequency Generator for impedance measurement and adjustment, as described in Technical Newsletter #2, provides a powerful measuring tool, most useful for wide bandwidths. When the impedance to be measured is part of a resonant system with a bandwidth less than about 10% of the center frequency, the delay line technique is less useful. The long line necessary to get the needed resolution has so much loss that measurement accuracy is reduced. The techniques described below are particularly useful with narrow-band circuits.**

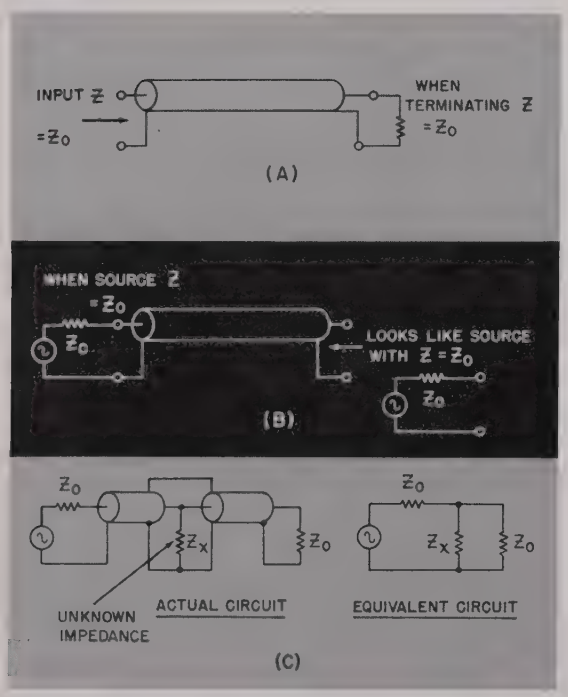



Figure 1—Measurement by Bridging Loss

### MEASUREMENT BY BRIDGING LOSS

It is basic in transmission line theory that the impedance seen looking into a section of line which is terminated at the far end in its characteristic impedance is the characteristic impedance—a pure resistance for low-loss lines. (See Figure 1A.) Similarly, the equivalent circuit seen looking back into a line fed from a matched source is a constant voltage in series with a resistor equal to $Z_0$ (Fig. 1B).

It follows that the circuit seen at any point along a matched system where the line from the source joins the line to the load is equivalent to a resistive source feeding a resistive load.

When a given impedance is "bridged," i.e. connected across an otherwise matched system, the loss of energy transferred to the load is the same as the loss that is calculated when bridging that same impedance across the load in the equivalent circuit (Figure 1C).

In the general case where the unknown impedance had both resistance and reactance, it would be necessary to know both the magnitude and phase angle of the loss introduced to determine the unknown constants. Where suitable equipment is available to measure both quantities, a chart has been worked out relating them to the constants of the bridging load (see "A Precise Sweep-Frequency Method of Vector Impedance Measurement" by D. A. Alsberg, Proc. I.R.E. Nov. 1951. PP 1393-1400).

In many practical situations it is known, from other considerations, that the unknown is either a pure resistance or a pure reactance. In these cases the magnitude of the unknown is determined by a simple attenuation measurement. The relationships can be calculated as shown in Appendix I; the results can be drawn up as nomographs relating bridging loss to attenuation. Two such charts have been made, and are illustrated on page 6 as Figure 9 (for systems with 50 ohm impedance) and Figure 10 (for 75 ohm systems).

## MEASUREMENT BY SERIES LOSS

"Bridging" is not the only way in which an unknown impedance can be connected into a matched system. It is also possible to open the center conductor at a junction and insert the unknown in series, as indicated in Figure 4.

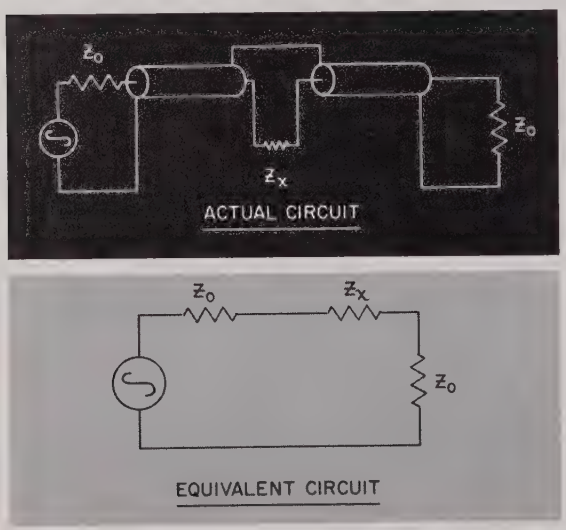



**Figure 2—Measurement by Series Loss**

The loss that is introduced is a measure of the impedance in this case also. For unknowns that are purely resistive or purely reactive, the loss can be calculated (see Appendix I, page 8) and the resulting relationship drawn up in the form of Nomographs: (See page 7, Figure 11 for Series Loss in 50 ohm systems; and Figure 12 for 75 ohm systems).

## TECHNIQUE

Certain precautions are essential if reasonable accuracy is to be attained:

1. Accurate impedance measurement depends on accurate attenuation measurement. The comparison technique, diagrammed in Figure 3, is recommended. In this set-up the loss in one branch, containing the unknown, is compared with the loss in another, containing an accurate variable attenuator. This procedure is described in more detail in Tech. Newsletter #1. The use of carefully matched fixed attenuator pads with 5 or 10 db loss at the points indicated in Figure 3 will guard against impedance mismatch in the associated equipment and increase measurement accuracy. The use of such pads is only possible where the unknown has moderate loss, or where the measuring set-up includes a high-gain rf amplifier.

2. Any convenient physical arrangement may be used for a test jig provided that the system characteristic impedance is maintained. Two connectors mounted on a metal sheet can be used, as illustrated in Figure 4. The dimensions of the sheet should be several times the spacing of the connectors to minimize stray couplings from the circuit under test to the rest of the universe. Where the unknown circuit to be tested is itself coaxial (as where a short piece of cable is being tested) it is most convenient to use a "Tee" connector to attach the unknown to the junction.

3. Harmonics or other spurious components in the output of the sweep-frequency generator can cause errors, particularly when the unknown gives an attenuation maximum at resonance. Figure 5 illustrates some techniques that can be used to minimize the effects of spurious signals. When the measurements are being made within an octave band, a low pass filter cutting off at the upper end of the octave can be used to reduce harmonic errors (Figure 5A). Where there is a choice in regards to the circuit configuration, the use of those that give attenuation minima at resonance will help to reduce the effects of harmonics (Figure 5B). Finally, a tuned amplifier or receiver of moderate bandwidth (e.g. a television or radar receiver) can be used to minimize harmonics as well as to increase the gain of the system (Figure 5C).

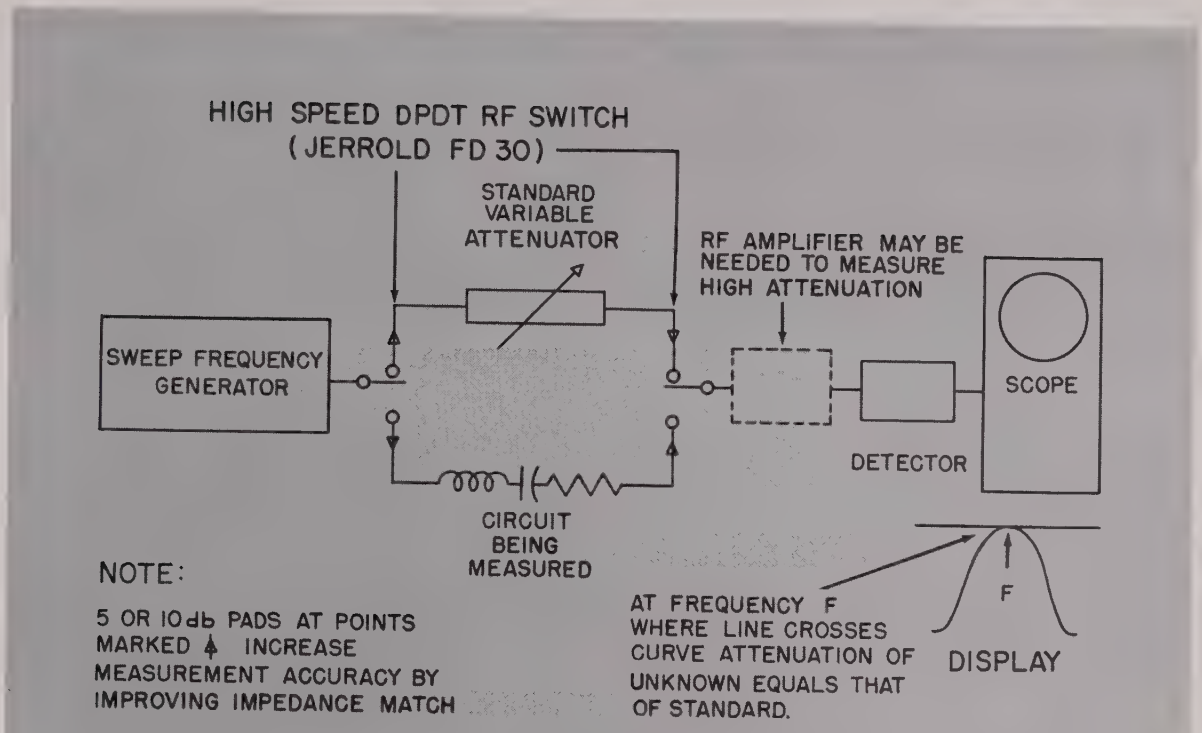
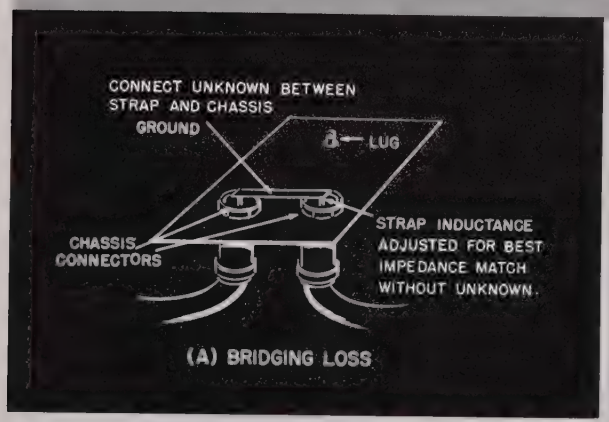



Figure 3—Technique for Accurate Attenuation Measurement



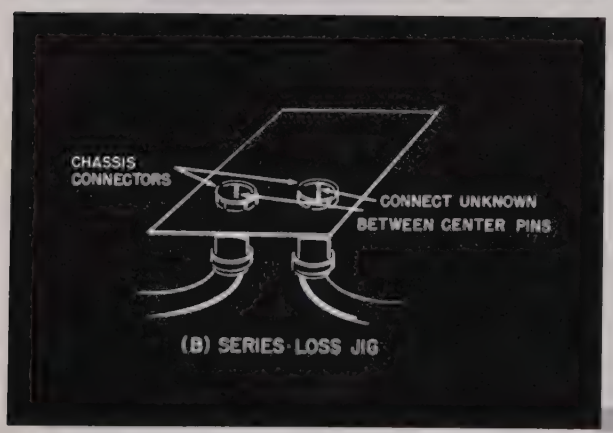




Figure 4—Test Jigs

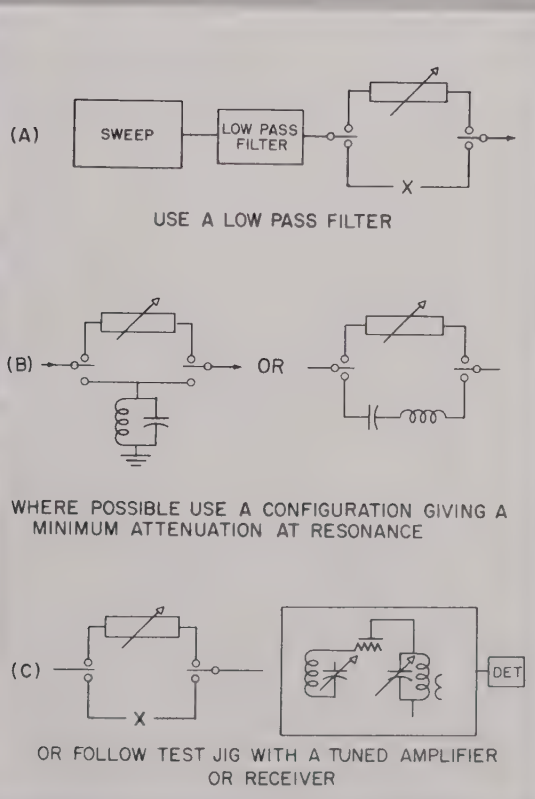
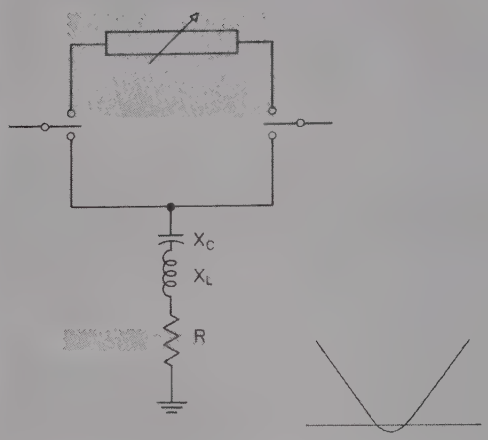



Figure 5—To Minimize Errors Due to Harmonics

## APPLICATIONS AND EXAMPLES OF RESULTS

1. **Measurement of Tuned Circuit Q** (See Figure 6). A convenient way to measure the unloaded Q of a resonant circuit is to determine its resonant resistance by the bridging or series loss method. The Q is then found by comparing this resistance with the reactance of either reactor at resonance, calculated from a known L or C value. The choice as to the use of the bridging or series connection, and series or parallel resonance is determined by the impedance level and physical configuration of the circuit being tested. As an example, a coil space-wound with 15 turns of #16 copper wire was connected in series with a 15 mmfd TCZ ceramic capacitor. Its bridging loss as a series resonant circuit connected across a 75 ohm system was 31 db at 35 MC. From the chart (Figure 10) this indicates a series resistance of 1.1 ohms. The reactance of 15 mmfd at 35 MC is 303 ohms, so the circuit Q was $\frac{303}{1.1}$ or 275.




$X_C = X_L$ AT RESONANCE.
MEASURE BRIDGING LOSS
AT RESONANCE, FIND R FROM
CHART.

$$Q = \frac{X_L}{R}$$

**Figure 6—Measuring Q**

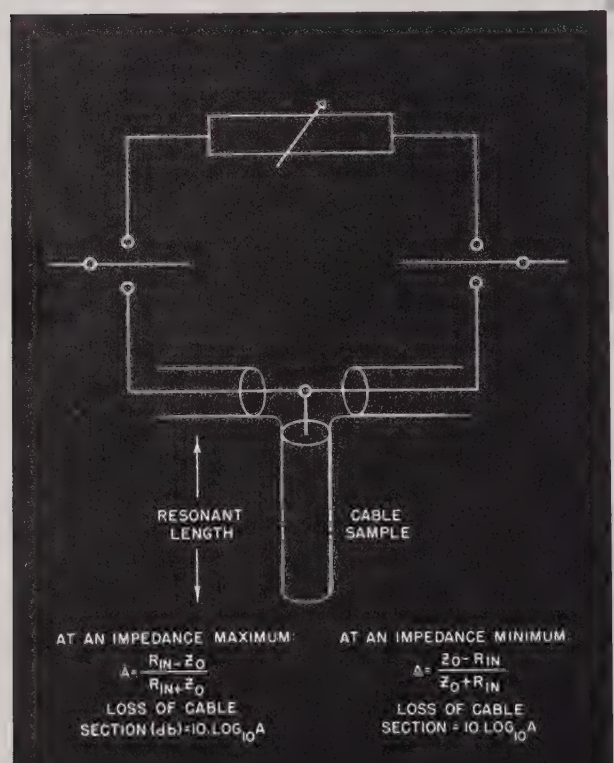

RESONANT LENGTH
CABLE SAMPLE
AT AN IMPEDANCE MAXIMUM
$A = \frac{R_{IN} - Z_O}{R_{IN} + Z_O}$
LOSS OF CABLE SECTION (db) = $10 \log_{10} A$
AT AN IMPEDANCE MINIMUM
$A = \frac{Z_O - R_{IN}}{Z_O + R_{IN}}$
LOSS OF CABLE SECTION = $10 \log_{10} A$

**Figure 7—Measuring Loss of Short Cable Section**

2. **Measurement of Attenuation of Short Cable Samples:** The input resistance seen looking into a short section of transmission line which is open or shorted at the far end is a function of the loss of this line section. Using a coaxial measuring system, the input resistance of a resonant section of coaxial line can be measured most conveniently by the bridging method. (See Figure 7.) The measured resistance is directly related to the attenuation of that line section, and can be used to calculate it (see Appendix II, page 8).

As an example, the input resistance of a 3 foot section of RG-6/U open at the far end was measured by the bridging method and found to give a bridging loss of 31 db at its quarter-wave resonant frequency of 160 MC. From the chart (Figure 10, page 7) this is found to correspond to a resistance of 1.1 ohms. This gives an attenuation for this sample of $10 \log_{10} \frac{75-1.1}{75+1.1}$ or 0.12 db; a loss of 4.0 db per 100 ft.

### 3. The Input Impedance of a Vacuum Tube:

Measuring the input impedance of a vacuum tube at high frequencies can be accomplished very simply by making this impedance part of a series resonant circuit by connecting an inductor in series with the grid terminal. Connecting the other end of the inductor as a bridging load across the measuring circuit allows rapid determination of the bridging loss, and thus of the total series resistance. (See Figure 8.) The measured resistance includes the losses of the inductor $R_L$ as well as those of the tube Rin). $R_L$ is small if a high Q inductor is chosen, but it can be measured quite accurately by replacing the tube with a high grade air trimmer set to give resonance at the same frequency, and measuring the bridging loss of the resulting circuit.

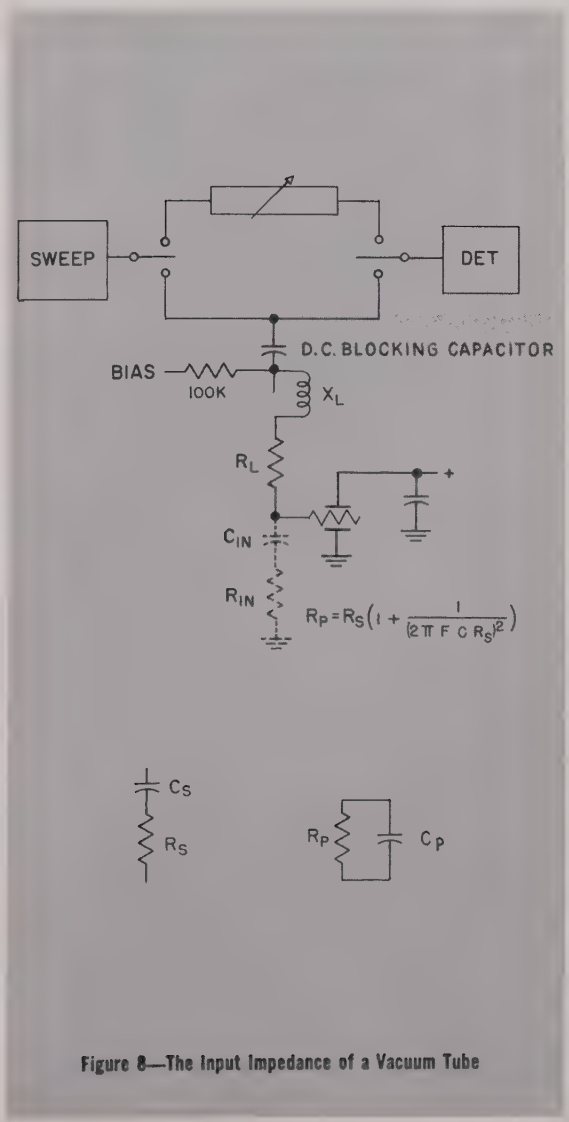



**Figure 8—The Input Impedance of a Vacuum Tube**

When measuring the input resistance of a tube by this method, it is notable that the measured resistance is only slightly affected by frequency. The results of measuring the input impedance of a 6CY5 tube with three different inductors are shown below.

| Inductance | 15.5 mh | 1.08 mh | 0.1 mh |
|---|---|---|---|
| Resonant Frequency | 14.15 MC | 54.3 MC | 188 MC |
| Bridging Loss | 12 | 14.3 | 13.3 |
| Series R | 12.5 | 9 | 10 |
| Coil R | 5.9 | 1.9 | 0.6 |
| Tube R | 6.6 | 7.1 | 9.4 |

It is commonly stated that the input resistance of a tube is approximately an inverse function of $f^2$. (See for example RCA Application Note AN-118 PP 8-9.) This statement is found to agree with the constant resistance described above when it is remembered that the R that is proportional to $\frac{1}{f^2}$ is the equivalent **parallel** resistance, while that which is approximately constant with frequency is the equivalent **series** resistance. Calling the former Rp and the latter Rs it can be easily shown that:

$$Rp = Rs\left(1 + \frac{1}{(2nfRsC)^2}\right)$$

If Rs is assumed constant, and Q is greater than 5 or so, this says that Rp is approximately equal to a constant divided by $f^2$.

Although both viewpoints are completely correct, it would seem more convenient to measure and think in terms of the relatively constant series resistive component, rather than the extremely variable parallel one.

# Bridging Loss Nomographs

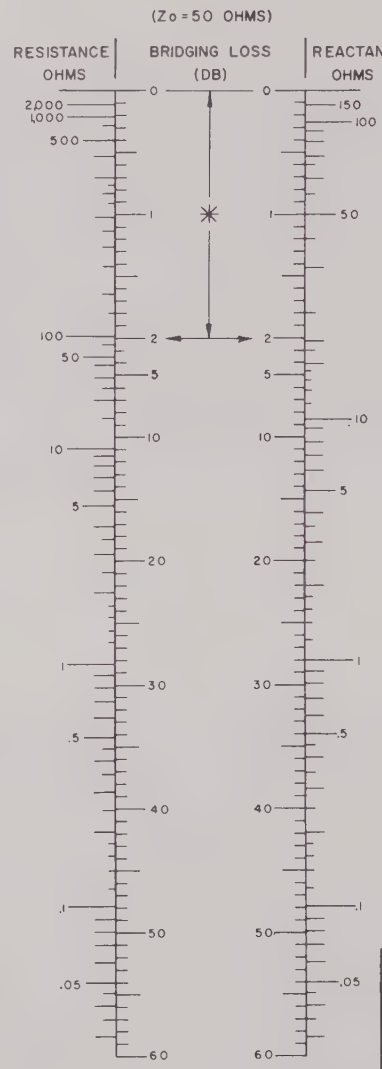
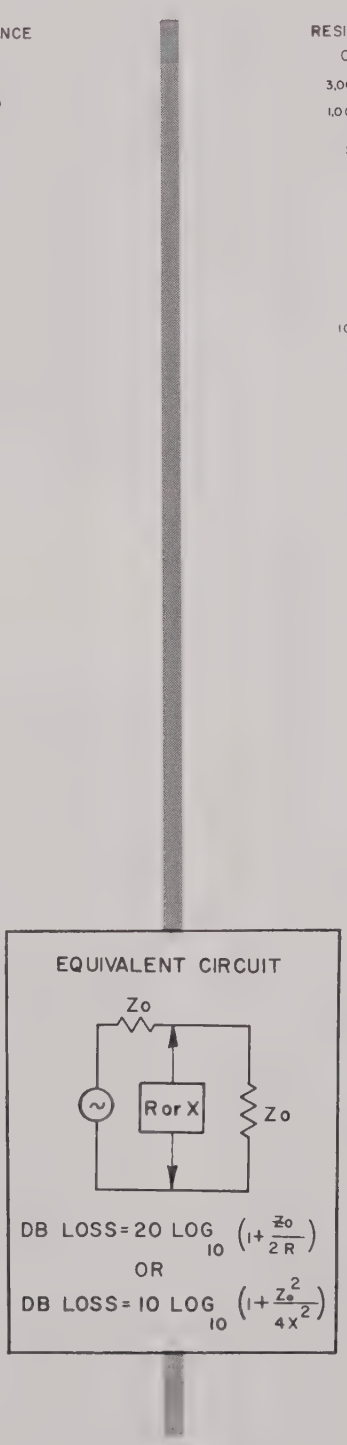



Figure 9




NOTE:
✳ EXPANDED SCALE 0-2 DB

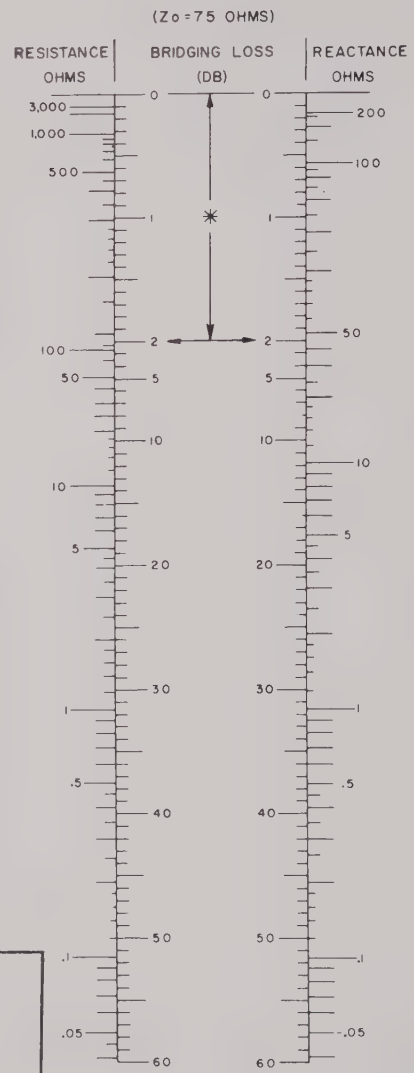



Figure 10

# Series Loss Nomographs

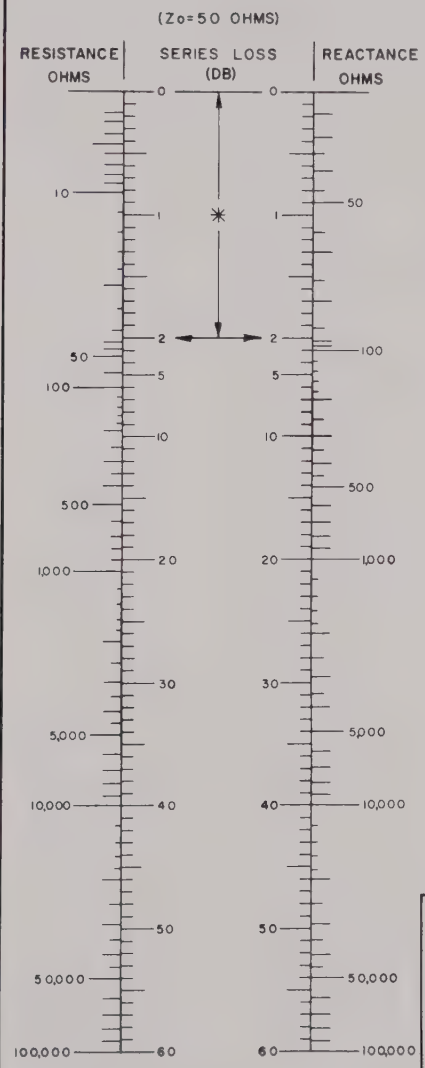
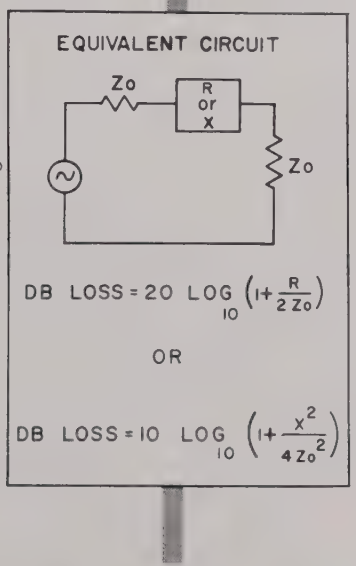



Figure 11




NOTE:
✱ EXPANDED SCALE 0–2 DB

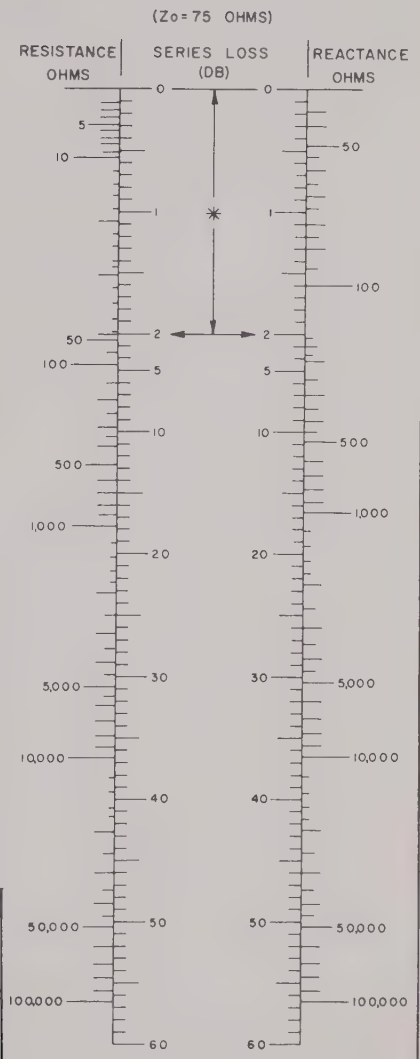
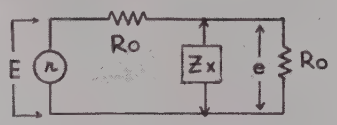



Figure 12

## APPENDIX I

**(a) The Calculation of Bridging Loss:**
**Equivalent circuit:**




**1. for Zx = a pure resistance "Rx"**

$$e = E\frac{\frac{Ro\,Rx}{Ro+Rx}}{Ro+\frac{Ro\,Rx}{Ro+Rx}} = E\frac{Rx}{Ro+2Rx}$$

$$e\ max = \frac{E}{2} \qquad \frac{e\ max}{e} = \frac{\frac{E/2}{E\,Rx}}{\frac{Ro+2Rx}{}} = \frac{Ro}{2Rx}+1$$

**insertion loss due to Rx =**

$$20\log_{10}\left|\frac{e\ max}{e}\right| = 20\log_{10}\left(\frac{Ro}{2Rx}+1\right)$$

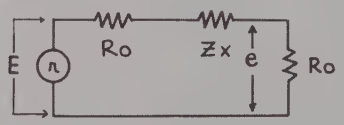
**2. for Zx = a pure reactance "jx"**

$$e = E\frac{\frac{j\,Rox}{Ro+jx}}{Ro+\frac{j\,Rox}{Ro+jx}} = E\frac{jx}{Ro+2jx}$$

$$\frac{e\ max}{e} = \frac{E/2}{\frac{E\,jx}{Ro+2jx}} \qquad \frac{Ro}{2jx}+1, \left|\frac{e\ max}{e}\right| = \sqrt{\left(\frac{Ro}{2x}\right)^2+1}$$

**insertion loss due to x =** $20\log_{10}\sqrt{\left(\frac{Ro}{2x}\right)^2+1}$

**(b) The Calculation of Series Loss:**
**Equivalent circuit:**




**1. for Zx = a pure resistance "Rx"**

$$e = E\frac{Ro}{2Ro+Rx}$$

**when Rx = o** $e\ max = E/2, \quad \frac{e\ max}{e} = \frac{E/2}{\frac{E\,Ro}{2Ro+Rx}} = \frac{Rx}{2Ro}+1$

**insertion loss due to Rx =** $20\log_{10}\left|\frac{e\ max}{e}\right| = 20\log_{10}\left(\frac{Rx}{2Ro}+1\right)$

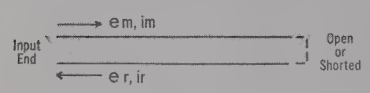
**2. for Zx = a pure reactance "jx"**

$$e = E\frac{Ro}{2Ro+jx} \qquad \frac{e\ max}{e} = \frac{jx}{2Ro}+1$$

**insertion loss due to x =** $20\log_{10}\left|\frac{e\ max}{e}\right| = 20\log_{10}\sqrt{\left(\frac{x}{2Ro}\right)^2+1}$

## APPENDIX II

**The Relation Between Input Resistance and Attenuation for a Resonant Length of Transmission Line**




When the far end is open or short-circuited, the reflected voltage or current wave at the input terminals is reduced, compared with the main wave by twice the attenuation of the line expressed in db (since the reflected wave makes one trip down and one trip back to reach the input terminals). Calling the round trip attenuation "a" (as a current or voltage ratio:

$$er = a\,em, ir = aim$$

**(a) At a frequency where the input impedance is a maximum**

$$R_{IN} = \frac{e\,max}{i\,min} = \frac{em+er}{im-ir} = \frac{em}{im}\left(\frac{1+a}{1-a}\right) = Zo\left(\frac{1+a}{1-a}\right)$$

$$Zo+a\,Zo = R_{IN}-a\,R_{IN}$$

$$\text{So} \qquad a = \frac{R_{IN}-Zo}{R_{IN}+Zo}$$

**the one-way attenuation of this section (db)** $= \frac{20\log_{10}a}{2}$

$$= 10\log_{10}\frac{R_{IN}-Zo}{R_{IN}+Zo}$$

**(b) At a frequency where the input impedance is a minimum**

$$R_{IN} = \frac{e\,min}{i\,max} = \frac{em-er}{im+ir} = \frac{em}{im} \times \frac{1-a}{1+a} = Zo\,\frac{1-a}{1+a}$$

$$R_{IN}+a\,R_{IN} = Zo-a\,Zo, \text{ so } a = \frac{Zo-R_{IN}}{Zo+R_{IN}}$$

**the one-way attenuation of this section (db) =**

$$10\log_{10}\frac{Zo-R_{IN}}{Zo+R_{IN}}$$

*Represented by*

**NOTE — Additional copies of this paper and engineering assistance on the application of Measurements By Comparison may be obtained by contacting the Industrial Products Division, Jerrold Electronics Corporation, 15th & Lehigh, Phila. 32, Pa. Phone: BAldwin 6-3456 — Ext. 211.**


EDITORIAL STAFF:
MANAGING EDITOR - C. C. Cooley, Jr.
TECHNICAL COPY EDITOR - A. Kirkaldy, Jr.
ART DIRECTOR - R. S. Reisner, Jr.

**Export: Rocke International, New York 16, N.Y. Cable ARLAB**
**Canada: Jerrold Electronics (Canada) Ltd., Toronto, Canada**

FORM NO. TN-3 BEM, JANUARY 1961 PRINTED IN U.S.A.

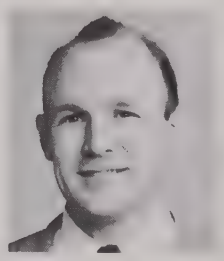
JERROLD

Technical Newsletter

Published by JERROLD ELECTRONICS CORPORATION, Philadelphia, Pa.
INDUSTRIAL PRODUCTS DIVISION

Volume 1, Number 4
COPYRIGHT © 1960 BY JERROLD ELECTRONICS CORP., PHILADELPHIA 32, PA.
August, 1960

# A Bridge Method of Sweep Frequency Impedance Measurement

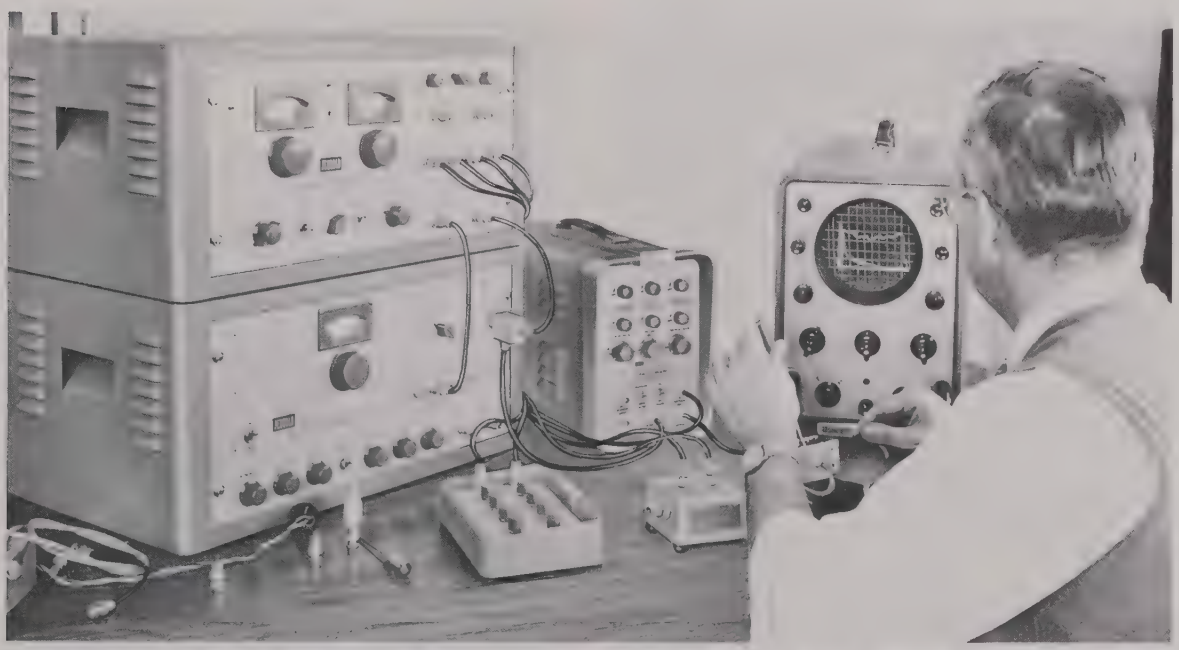


by KEN SIMONS, Chief Engineer, Jerrold Electronics Corp.

The following is an extract of a speech given by Mr. Simons at the First Annual Eastern Engineering Seminar, May 5, 1960, Hotel Bradford, Boston, Massachusetts.

## SUMMARY:

**A matched bridge circuit is described providing oscilloscopic display and measurement of return loss vs. frequency. Units have been constructed allowing measurement of 50 db return loss over a 50 to 1 frequency range (4-200 MC). The technique is useful from low audio frequencies up to several hundred megacycles.**



A typical test set-up utilizing Jerrold Models—900-A Sweep Frequency Generator, VC-12 Voltage Comparator, CM-6 Marker Generator, AV-75 Variable Attenuator and KSB-75 ohm Bridge; to show an optimum termination, 4 mcs to 100 mcs, with a 50 db reference.

# FUNDAMENTALS:

Since high frequency measurements frequently require unavoidable lengths of transmission line, high frequency measuring techniques usually involve an application of transmission line principles. The bridge circuit to be described is no exception. Understanding its operation requires a clear picture of the concepts of reflection coefficient and return loss.

Consider an AC generator with internal impedance Ro and a voltage E as illustrated in Figure 1 (a). When the output

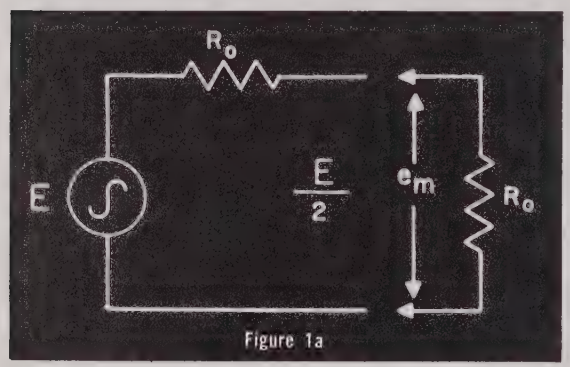


Figure 1a

of this generator is connected to a load having an impedance equal to Ro, the output voltage is $\frac{E}{2}$. In the particular case where this load is the input impedance of a section of transmission line having a characteristic impedance Ro, it is convenient to consider the input voltage as consisting of two components—a main voltage wave ($e_m$) which travels down the line away from the generator, and a reflected wave ($e_r$) which travels back from the far end towards the generator. When the transmission line section is terminated in Ro, the reflected wave is zero and the input voltage is simply $e_m$ Figure 1 (b).

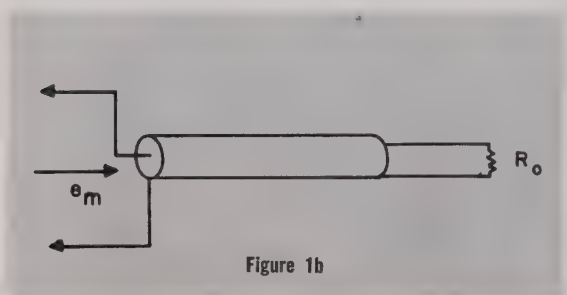


Figure 1b

This terminology can be applied even when there is no transmission line in the circuit. It may correctly be said in reference to Figure 1 (a) that the voltage across the load Ro is $e_m$.

In the case where the generator feeds a section of transmission line terminated in some impedance other than Ro, the input voltage can be considered to be the sum of $e_m$ and $e_r$ (Figure 1c). Similarly the voltage across any load

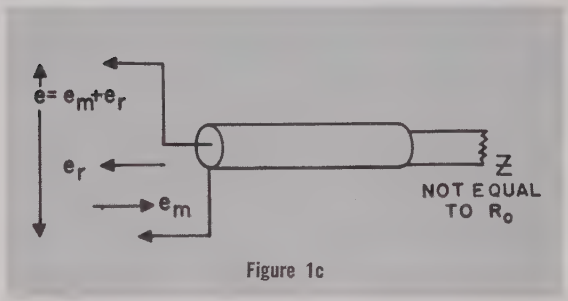


Figure 1c

connected to a generator can be considered to be the sum of a main wave and a reflected wave (Figure 1d).

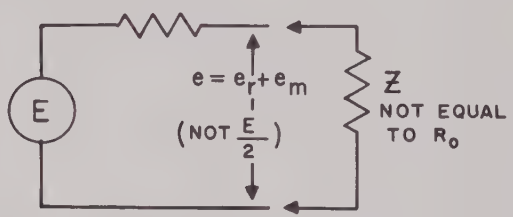
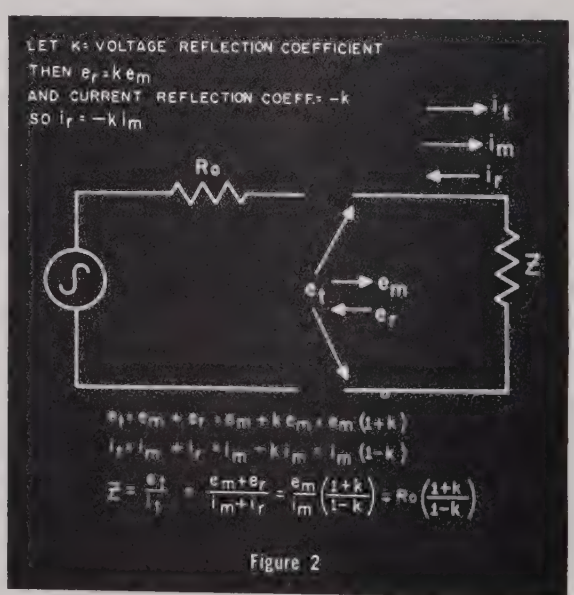


Figure 1d

A convenient way of describing the relation between the main and reflected waves is to use the reflection coefficient. Figure 2 illustrates a generator feeding a load having any impedance "Z". The voltage and current across this load


LET K= VOLTAGE REFLECTION COEFFICIENT
THEN er=k em
AND CURRENT REFLECTION COEFF.= -k
SO ir = -k im
it
im
ir
Ro
Z
et
em
er
et = em + er = em + k em = em (1+k)
it = im + ir = im - k im = im (1-k)
Z = et/it = (em+er)/(im+ir) = em/im ((1+k)/(1-k)) = Ro ((1+k)/(1-k))
Figure 2

are expressed in terms of their main and reflected components. The quantity k is defined as the ratio of the reflected voltage $e_r$ to the main voltage $e_m$. The current reflection coefficient is numerically equal to the voltage reflection coefficient and has the opposite sign, so that the relationships indicated in Figure 2 hold true. A significant point is that the impedance of the load can be fully described by the internal impedance of the generator and the voltage reflection coefficient. The relation is:

$$Z = Ro \frac{1 + k}{1 - k}$$

Note that k is a complex number involving the relation between not only the magnitudes of the main wave and reflected wave, but also their phase angles.

A further convenience is achieved by expressing the reflection coefficient logarithmically. The "Return Loss" is defined as 20 $\log_{10}$ k. Thus, it is the ratio of $e_r$ to $e_m$ expressed in decibels. The interrelation between Return Loss, Reflection Coefficient and VSWR is indicated on the Nomograph of Figure 3.

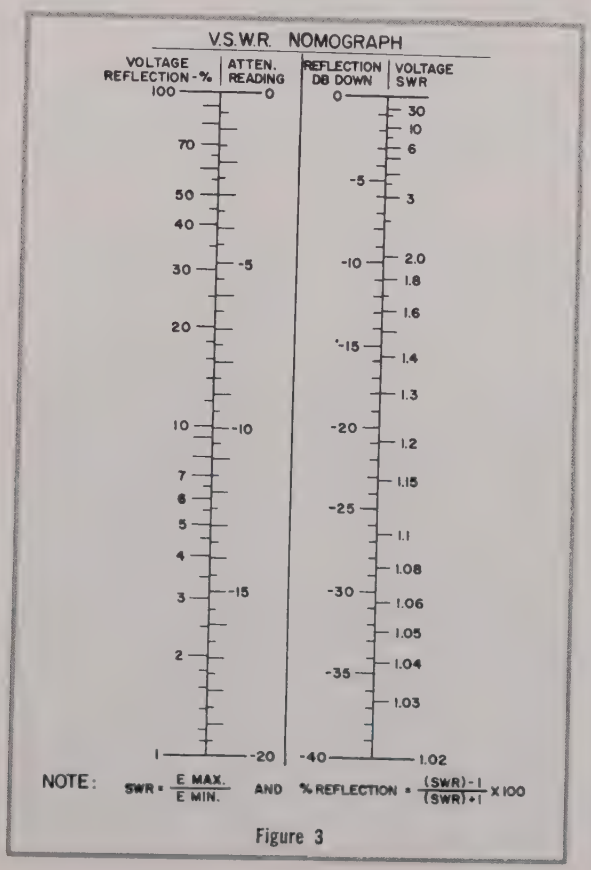

V.S.W.R. NOMOGRAPH
VOLTAGE REFLECTION - %
ATTEN. READING
REFLECTION DB DOWN
VOLTAGE SWR
NOTE: SWR = E MAX. / E MIN. AND % REFLECTION = (SWR)-1 / (SWR)+1 X 100

Figure 3

# THE BRIDGE PRINCIPLE:

A convenient way of measuring an unknown impedance would be to separate the reflected wave from the main wave and to measure its relative magnitude and phase angle. This is commonly done in high-frequency and microwave testing by the use of a directional coupler. Another approach is the use of a bridge.

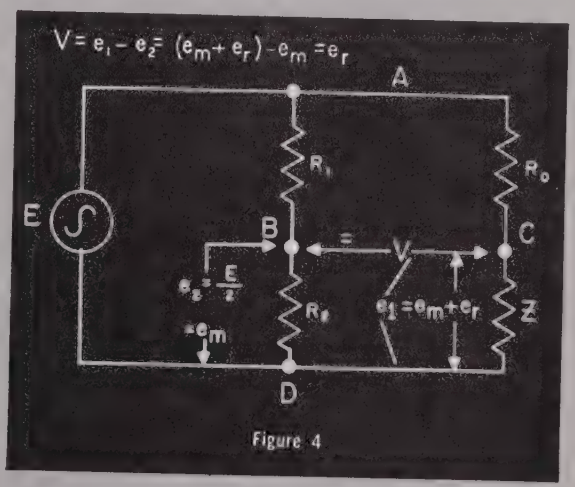



Figure 4

Figure 4 illustrates how a bridge can provide a voltage equal to the reflected wave. A generator having zero internal impedance feeds a load Z through a resistor Ro. The voltage $e_r$ across Z is equal by definition to $e_m$ plus $e_r$. If a second divider consisting of two equal resistances $R_1$ and $R_2$ is connected across the voltage source, the voltage $e_2$ across the lower resistor will be equal to $\frac{E}{2}$ which is the same as $e_m$. If the voltage between the center points of the two dividers is measured, it is found to be the difference between $e_m$ plus $e_r$ and $e_m$, which is $e_r$. Thus, the output voltage "V" of this network is identical to the reflected voltage component from the impedance Z.

# THE TERMINATED BRIDGE:

Unfortunately, this simple circuit involves the use of a voltage source having zero internal impedance and a voltage measuring device having infinite impedance which measures the difference of voltage between two points, neither of them grounded. As a practical matter, these conditions are difficult to approach.

Fortunately, an alternative configuration is available. The terminated bridge illustrated in Figure 5 has several characteristics which make it especially suitable for this application. When all six resistances are equal, the impedance seen looking into the network at the terminals left after disconnecting any one of the resistances is always equal to Ro. Thus, for example, in Figure 5 the impedance seen between A and D with #1 resistor removed is equal to Ro. When the resistors are equal to the characteristic impedance of a particular coaxial cable, the bridge has matched input and output impedances.

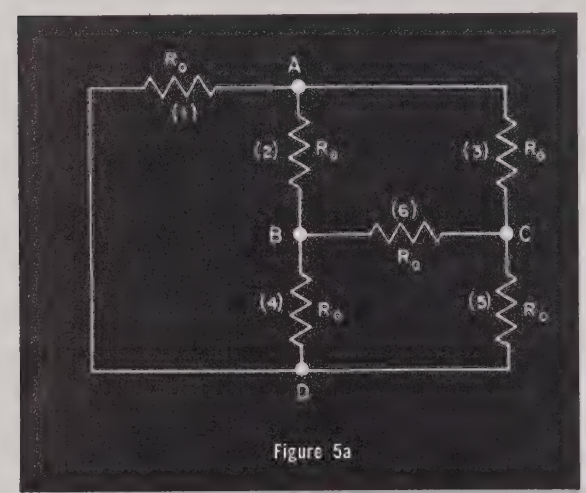



Figure 5a

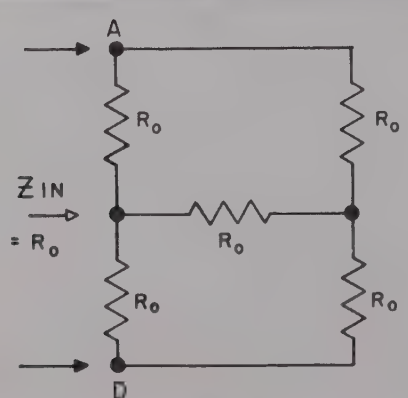
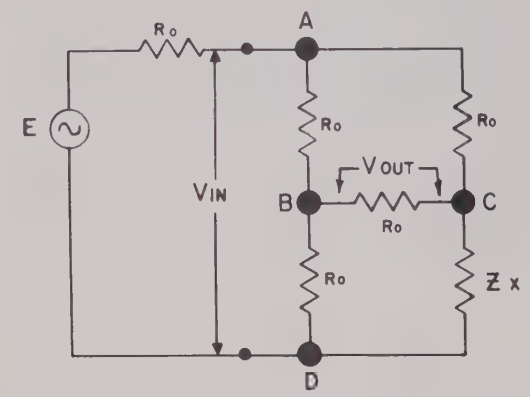



Figure 5b

The terminated bridge has the further advantage that it provides an output voltage precisely equal in magnitude and phase to a constant times the reflection coefficient of the unknown. This is illustrated in Figure 6 which states




TRANSMISSION LOSS $\left(20 \log_{10} \frac{V_{IN}}{V_{OUT}}\right) =$

RETURN LOSS OF $Z_X$ + 12 db $\left(= 20 \log_{10} \frac{1}{4} \frac{Z_X - R_o}{Z_X + R_o}\right)$

Figure 6

that the transmission loss from Vin, the input terminals of the bridge, to Vout is equal to the return loss of Zx (referred to Ro) plus 12 db. This relation is fully derived in Appendix I. It may be understood better by referring to Figures 7 and 8. Figure 7 shows that the input voltage to a matched load connected between C and D ($e_m$) is equal to the bridge input Vin reduced by 6 db.

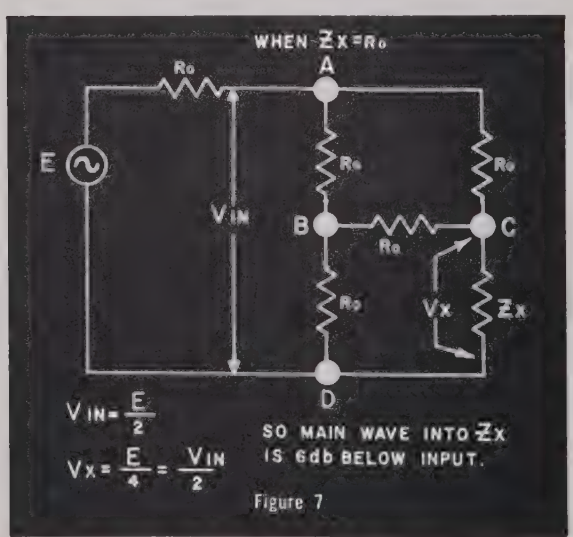



Figure 7

In Figure 8 the bridge is redrawn to show the voltage division experienced by the return wave from Zx. It comes back into terminals C and D and is attenuated 6 db before coming out of terminals B and C.

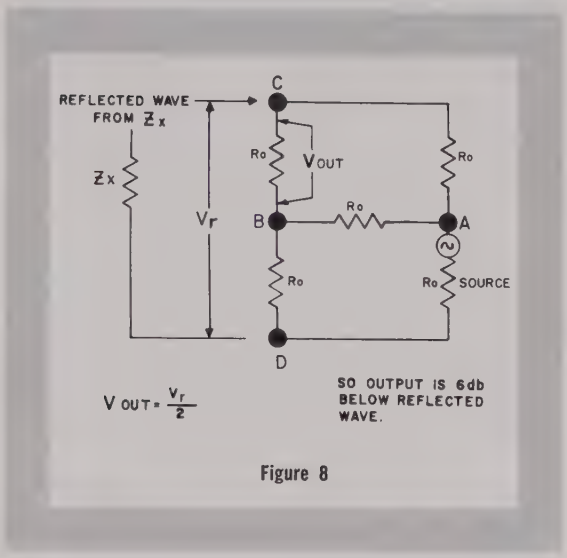



Figure 8

As compared with the simple bridge circuit illustrated in Figure 4, the matched bridge has the advantage that the input and output impedances match connecting coaxial cables so that wide band frequency performance can be obtained. The problem of providing a differential voltage measurement between two ungrounded terminals is still present. To build a matched bridge practically, it is necessary to provide a transformer having balanced input terminals as illustrated in Figure 9. It has been found

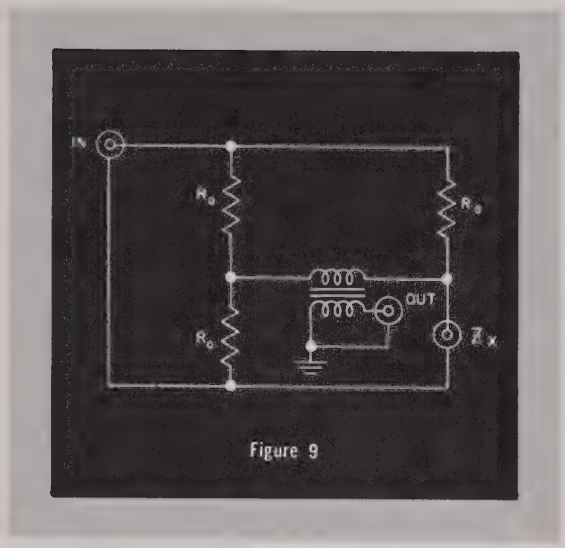



Figure 9

possible to build such a transformer with a very wide range of frequency response. The resulting bridge is conveniently connected to provide a calibrated sweep display as illustrated in Figure 10. A high-speed coaxial switch (Jerrold Model FD-30) is connected so that in the "up" position it connects the output of the sweep through a 12 db pad and a standard variable attenuator to the input of a wide band amplifier. The amplifier is followed by a detector and scope and displays a curve representing loss vs. frequency. In the "down" position the switches insert the bridge in place of the attenuators. Since the loss of the bridge is equal to the return loss of the unknown plus 12 db, the scope will show a plot of return loss vs. frequency with a reference line corresponding to the setting of the standard variable attenuator.

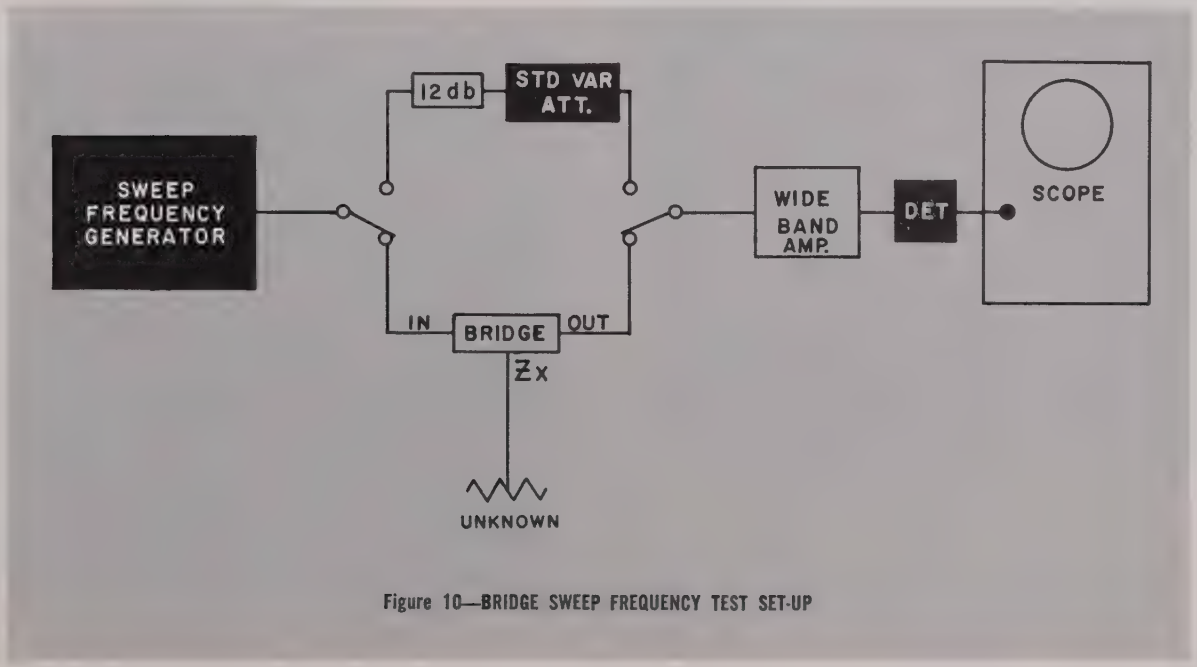



Figure 10—BRIDGE SWEEP FREQUENCY TEST SET-UP

# EXAMPLES OF USE

The performance of a particular bridge is illustrated in Figures 11 to 16 inclusive. This bridge was designed for a characteristic impedance of 75 ohms and the test circuit covered a frequency range from 4 to 100 MC. Figure 11 illustrates the relation between the reference line and the bridge response with the unknown terminal of the bridge open-circuited. The bridge indicates 0 db return loss within about ½ db from 4 to 100 MC.

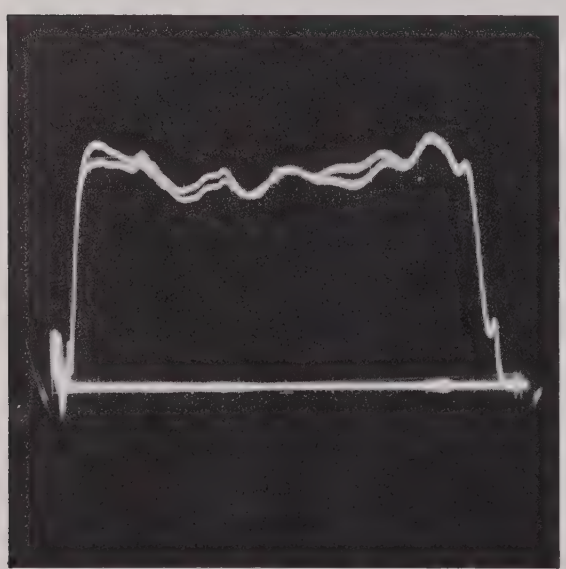


**Figure 11—BRIDGE RESPONSE**
**4-100 mc, Open Circuit**

Figure 12 shows the same situation with the unknown terminal short-circuited, again indicating 0 db return loss with approximately ½ db error.

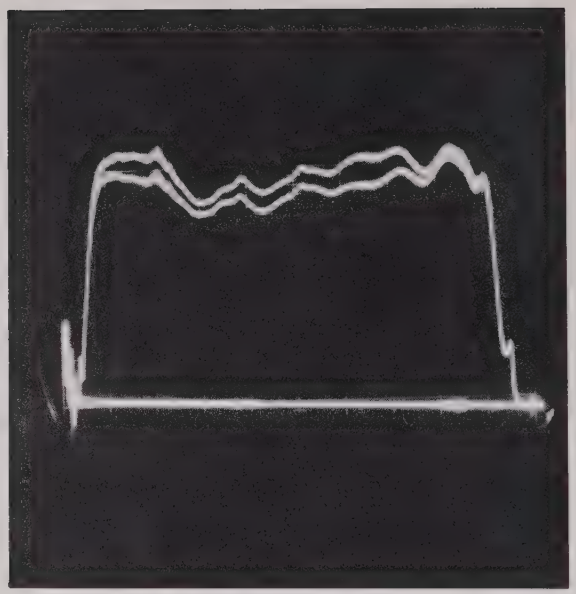


**Figure 12—BRIDGE RESPONSE**
**4-100 mc, Short Circuit**

To obtain Figure 13, the X terminal of the bridge was terminated as accurately as possible. In this figure the reference line represents a return loss of 50 db and it is seen that the loss through the bridge indicates a return loss substantially better than 50 db across this entire frequency range. A 50 db return loss corresponds to a VSWR of 1.06 so that it is apparent that this technique is capable of precision comparable with the best that can be obtained with other high frequency measuring techniques.

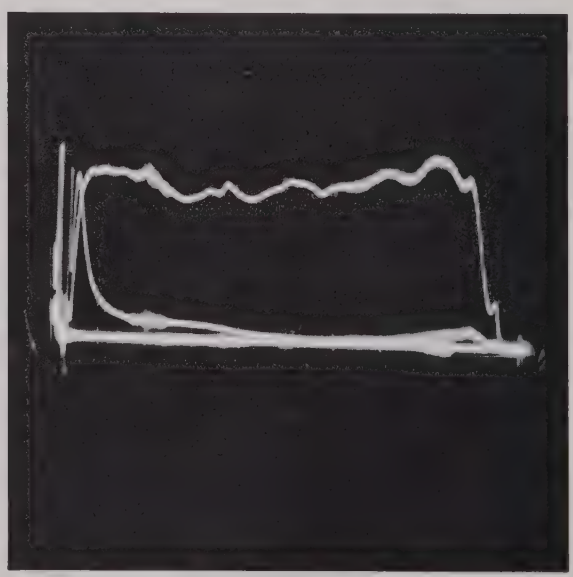


**Figure 13—BRIDGE RESPONSE**
**4-100 mc, Optimum Termination, 50 db Reference**

Although this method is extremely **precise,** i.e., it can compare an unknown with a standard and show deviations of considerably less than 1%, its **accuracy** is entirely dependent on the accuracy of the standard impedance against which the bridge is tested. Where suitable standards are available, its accuracy can be nearly as good as its precision.

Figure 14 illustrates the agreement between the reference line and the bridge loss when the bridge was first adjusted for balance with a 75 ohm resistor. A resistor was then substituted whose DC resistance was accurately adjusted to 70 ohms. A 70 ohm resistor in a 75 ohm system gives a VSWR of 1.07 or a return loss of 29.2 db. The reference line on this trace was 30 db.

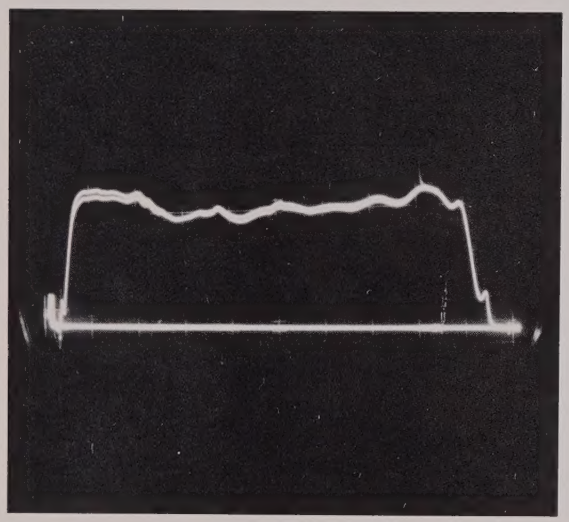


**Figure 14—BRIDGE RESPONSE**
**4-100 mc, 70 ohm Termination, 30 db Reference**

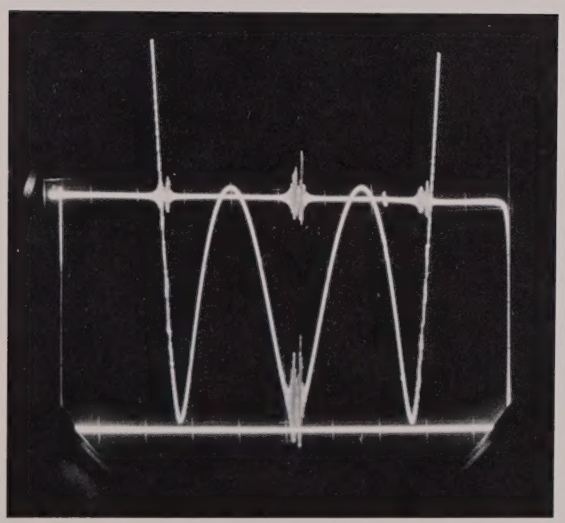


**Figure 15 —BRIDGE RESPONSE**
**Markers 100 mc ±1.5 mc, 10 db Reference**

The delay line method of sweep-frequency impedance measurement (as described in Technical Newsletter #2) provides a convenient method of measuring high frequency impedances with a minimum of equipment. Practical characteristics of delay lines limit the usefulness of this method to relatively wide frequency bands. The bridge technique has no such limitation. It is equally useful for extremely narrow-band measurements or extremely wide-band measurements.

Figure 15 shows the return loss measurement of a triple-tuned filter having a bandwidth of 1.5 MC and a maximum return loss in the pass band of approximately 10 db. Notice the simplicity of this presentation and the ease with which an operator could adjust such a filter to meet given frequency and return loss specifications.

Figure 16 illustrates an extremely wide band measurement. This shows the return loss of a piece of high-grade coaxial cable over the frequency range 4 to 100 MC with a 40 db reference. The bridge provides a way of testing this complicated function with a high degree of speed and accuracy.

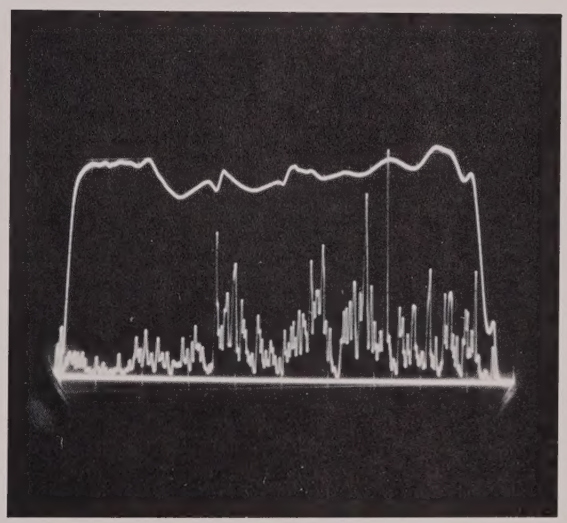
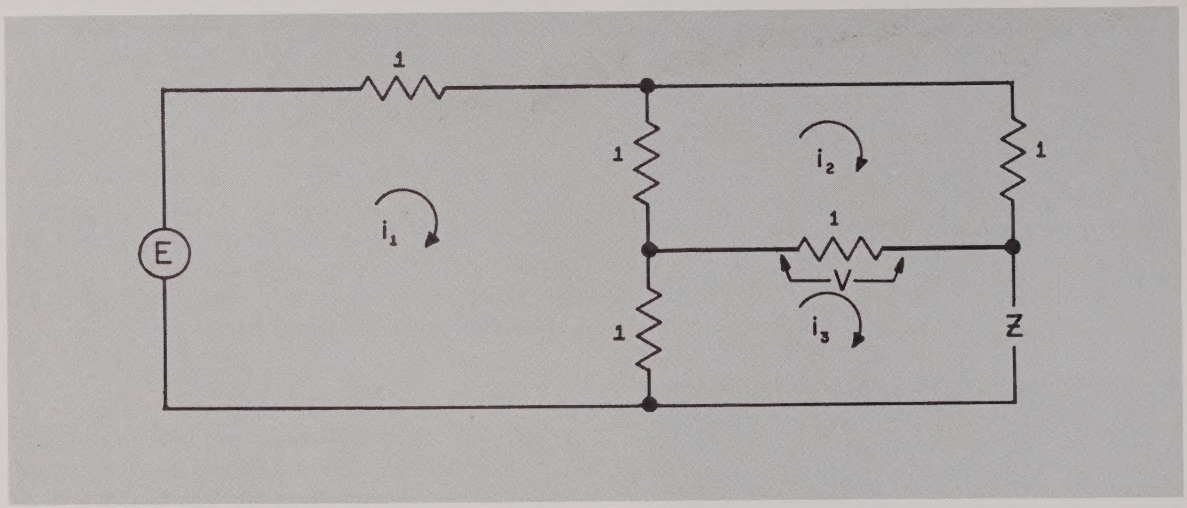


**Figure 16—BRIDGE RESPONSE**
**4-100 mc, Input of 1000' Piece High Grade Coax., 40 db Reference**

The sweep frequency technique has for many years provided a convenient way of observing the frequency variation of gain or loss. More recently it has become possible to measure these quantities simultaneously with the sweep presentation. (See Technical Newsletter #1.) The wide-band return-loss bridge now provides the same degree of convenience in viewing and measuring the frequency variation of return loss and opens a new area of speed and convenience in Laboratory and Production testing of high frequency networks.

# APPENDIX I




① $\ldots\ldots\ldots\ldots\ldots\ldots E = 3i_1 - i_2 - i_3$

② $\ldots\ldots\ldots\ldots\ldots\ldots 0 = i_1 + 3i_2 - i_3$

③ $\ldots\ldots\ldots\ldots\ldots\ldots 0 = -i_1 - i_2 + (2 + Z)\, i_3$

④ $= ② - ③ \ldots\ldots\ldots\ldots 0 = 4i_2 - (3 + Z)\, i_3$

⑤ $= ① + (3 \times ②) \ldots\ldots E = 8i_2 - 4i_3$

⑥ *Solving ④ for* $i_2 \ldots\ldots i_2 = \dfrac{Z + 2}{4} i_3$

⑦ *Substituting in ⑤* $\ldots\ldots E = \left[ 8\left(\dfrac{3 + Z}{4}\right) - 4 \right] i_3 = 2\,(Z + 1)\, i_3, \qquad i_3 = \dfrac{E}{2\,(Z + 1)}$

⑧ *Substituting in ⑥* $\ldots\ldots i_2 = \dfrac{Z + 3}{4} \cdot \dfrac{E}{2\,(Z + 1)} = \dfrac{E}{8} \cdot \dfrac{Z + 3}{Z + 1}$

⑨ $\ldots\ldots\ldots\ldots\ldots\ldots V = 1\,(i_2 - i_3) = E \dfrac{1}{8} \cdot \dfrac{Z + 3}{Z + 1} - \dfrac{1}{2\,(Z + 1)}$

*So:* $\boxed{V = \dfrac{E}{8} \cdot \dfrac{Z - 1}{Z + 1}}$

*Represented by*

**NOTE — Additional copies of this paper and engineering assistance on the application of Measurements By Comparison may be obtained by contacting the Industrial Products Division, Jerrold Electronics Corporation, 15th & Lehigh, Phila. 32, Pa. Phone: BAldwin 6-3456 — Ext. 211.**

EDITORIAL STAFF:
MANAGING EDITOR - C. C. Cooley, Jr.
TECHNICAL COPY EDITOR - A. Kirkaldy, Jr.
ART DIRECTOR - R. S. Reisner, Jr.

**Export: Rocke International, New York 16, N.Y. Cable ARLAB**
**Canada: Jerrold Electronics (Canada) Ltd., Toronto, Canada**

FORM NO. TN-4 BEM, JANUARY 1961 PRINTED IN U.S.A.

JERROLD®


**Jerrold Electronics Corporation**

INDUSTRIAL PRODUCTS DIVISION

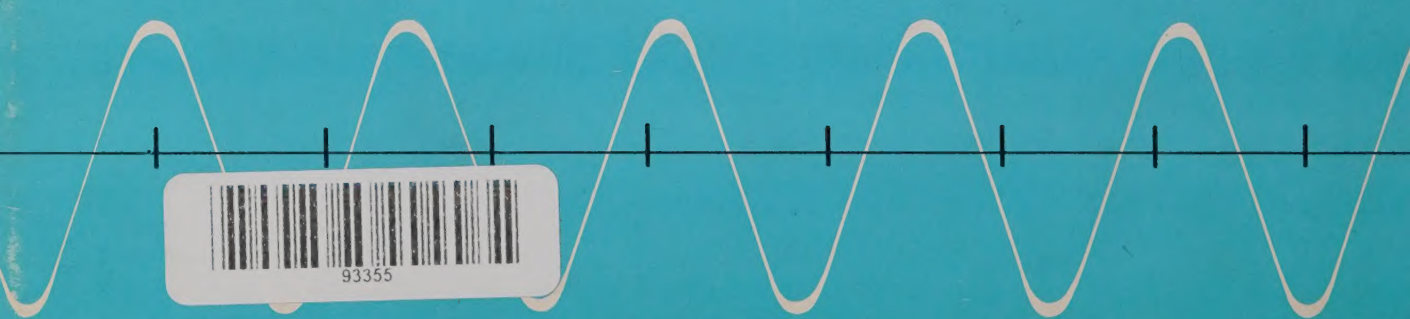
15th and Lehigh Avenue • Philadelphia 32, Pa.

BAldwin 6-3456


*REPRESENTED BY*

**EXPORT:** Rocke International, New York 16, N. Y., Cable ARLAB

**CANADA:** Jerrold Electronics (Canada) Ltd., Toronto, Canada




93355

Printed in U.S.A.